WIRELESS ENGINEER

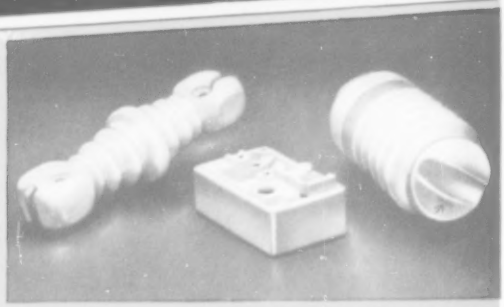
THE JOURNAL OF RADIO RESEARCH & PROGRESS

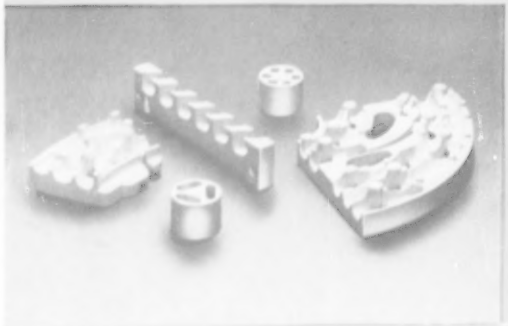
NOVEMBER 1956

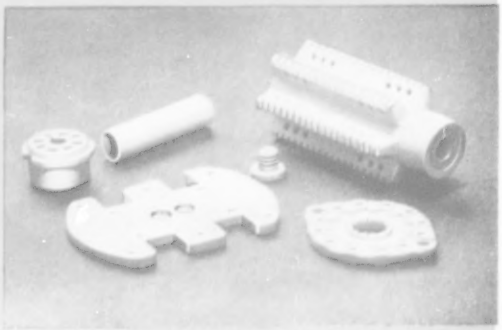
Vol. 33 No. 11 · THREE SHILLINGS AND SIXPENCE

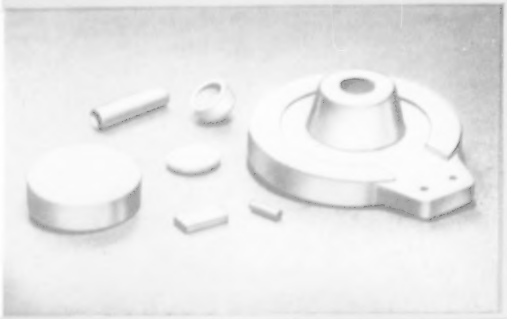
Bullers CERAMICS
FOR INDUSTRY

High quality material and dimensional precision are attributes of Bullers die-pressed products.
Prompt delivery at competitive prices.









We specialise in the manufacture of

PORCELAIN
for general insulation

REFRACTORIES
for high-temperature insulation

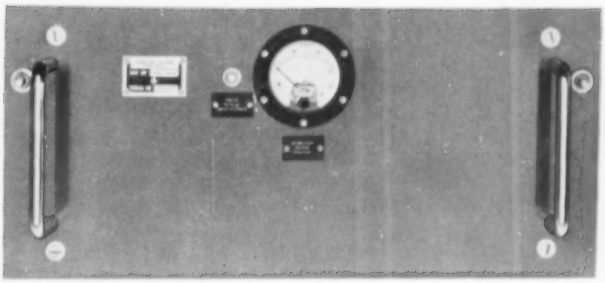
FREQUELEX
for high-frequency insulation

PERMALEX & TEMPLEX
for capacitors

BULLERS LIMITED
MILTON · STOKE-ON-TRENT · STAFFS
Phone: Stoke-on-Trent 21381 (5 lines) Telegrams & Cables: Bullers, Stoke-on-Trent
Ironworks: TIPTON, STAFFS. Phone: Tipton 1691
London Office: 6 LAURENCE POUNTNEY HILL, E.C.4 Phone: MANsion House 9971

BL130

SPECIAL A-C AUTOMATIC VOLTAGE STABILIZER



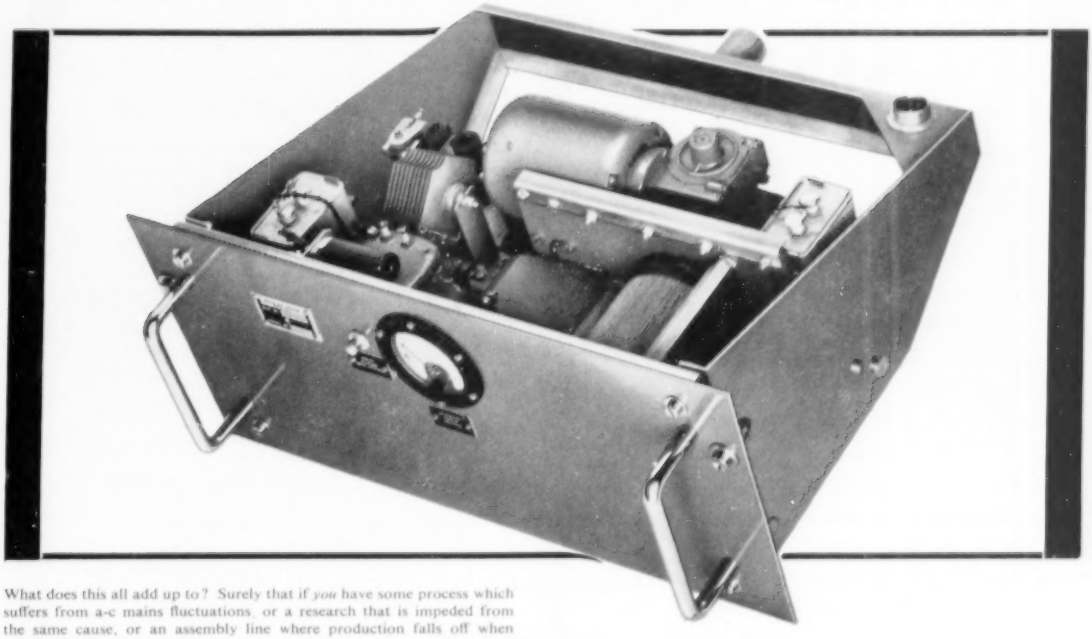
TYPE 8571 A.M. REF. No. 10D/20161. C. L. LTD. CAT. No. BMVR-2750-58-S

We are very proud to announce that some months ago, after the most exhaustive trials, a special relay-rack mounted and metered model of our regular Model BMVR-2750 A-C Automatic Voltage Stabilizer, fully tropicalized to Service requirements, was accepted for incorporation in *Radar Beacon Installations* as manufactured by Murphy Radio Ltd., for supply to the Royal Air Force.

Approaching one hundred are being supplied. Where only the best and most reliable equipment of this description is needed, and where failure might have the direst and most far-reaching consequences "play safe" and send us your enquiry. This is commonsense when the best costs no more, and is backed by stabilizer experience going back to pre-war days.

Some recent important Automatic Voltage Stabilizer Contracts received include sixty-six for Air Ministry, for supply to all Meteorological Offices (33 Type BAVR-1000 and 33 Type BMVR-7000-a) all delivered, and all accepted after very rigid inspection. Also forty-five specially mounted Type BMVR-2750-A (2.75 kVA) for supply to Air Ministry for use on another radar chain. Over 100 Type BMVR-1725 (1·725 kVA) have now been supplied to U.K.A.E.A.

Gillette Industries purchased a sample BAVR-1000 (1 kVA, instantaneous response, degree of stabilization 0·15%), and after trial ordered twenty-three more. This list of *recent* successes could be extended to fill a substantial proportion of this advertisement space.



What does this all add up to? Surely that if *you* have some process which suffers from a-c mains fluctuations, or a research that is impeded from the same cause, or an assembly line where production falls off when voltage is low you should examine the possibilities of our very considerable range of A-C Automatic Voltage Regulators. Single Phase models range from as low as £25.16.0 (ASR-1150, 1·15 kVA) to £238.13.0 (BMVR-25000-C, 29·00 kVA). So why delay? Please let us have your enquiry or problem, giving as much data as possible, addressing our nearest works.

For complete information on our entire line of Stabilizers and Regulators, of which we have *sixty-six distinct models*, please request a copy of our 20-page Catalogue Supplement Ref. V-549-S, together with our special publication VSP-56/16.

At the same time remind us not to forget to include our latest 12-page 3-Colour "VARIAC" Voltage-Regulating Transformer brochure, C.L.L. Form 424-UK. (Because in many cases it is of course possible to use manual control, which is considerably cheaper than automatic control: we do not want you to spend more than is necessary!)

Claude Lyons Ltd.

76 OLDHALL STREET · LIVERPOOL 3 · LANCS · VALLEY WORKS · WARE RD · HODDESDON · HERTS

Telephone: Central 4641/2 Tel.: Hoddesdon 3007 (4 lines)

Available for the first time

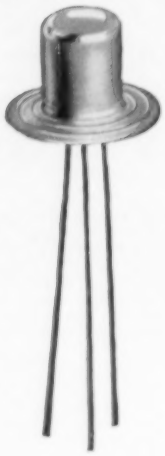
a complete range of

transistors

—for the radio

and electronics

industry



INTRODUCING

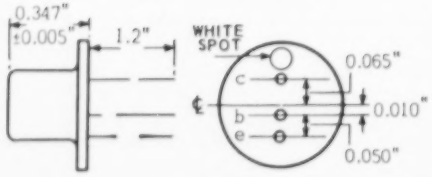
the New Metal Clad "Top Hat" TRANSISTORS

★ Hermetically sealed, welded metal-to-metal container insures against moisture penetration even in high humidity at high ambient temperatures.

★ Revolutionary new all-metal construction gives far greater heat dissipation.

★ A complete range of basic samples with mean average characteristics are available, including matched pairs for push-pull output, for Design Engineers in the radio and electronics industry.

TYPE	APPLICATION
XA101	I.F. Amplifier
XA102	Frequency Changer/Oscillator (cut-off frequency 7 Mc/s)
XB102	Intermediate L.F. Stage
XB103	Intermediate L.F. Stage
XC100 Series	Output Stage, including matched pairs for push-pull



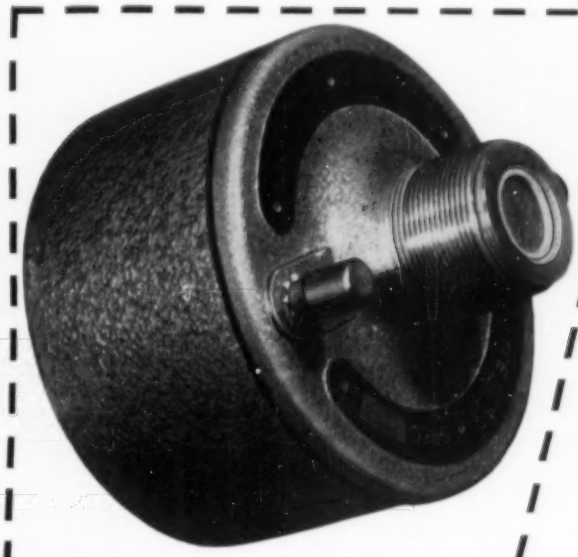
Send for range of Basic Samples and Technical Data.

EDISWAN
MAZDA

RV29

RADIO DIVISION

THE EDISON SWAN ELECTRIC CO. LTD., 155 CHARING CROSS ROAD, LONDON, W.C.2
Telephone: Gerrard 8660 *Member of the A.E.I. Group of Companies* Telegrams: Ediswan, Westcent, London



DC 10
PRESSURE UNIT

SPECIFICATION

Power Handling Capacity	10 watts peak
Voice Coil Impedance	15 ohms
Frequency Response	120–9000 c.p.s.
Flux Density	12,000 gauss
Pole Piece	1·5″ diameter

Price £6 10s. 0d.

DIMENSIONS

	DC 10 (without line trs.)	DC 12 (with line trs.)
Diameter	4⅛″	4⅛″
Length	4¼″	6⅜″
Weight	4¼lb.	5¼lb.

Specially designed and developed to meet the need of the P.A. Engineer requiring a compact, efficient unit combining good tone with average handling capacity, at a price that will make the "small" installation a profitable proposition.

High Sensitivity Heavy cross-sectioned cup with latest anisotropic alloy CP magnet.

Phase Equalizing Throat One-piece zinc based alloy die-casting.

Self-Centring Diaphragm Assembly Can be changed in the field without special tools or soldering iron in 1½ mins.

Spring Loaded Terminals Ensure quick and positive line termination.

Weatherproof Totally enclosed, ensuring protection when in exposed position, watertight gland cable entry.

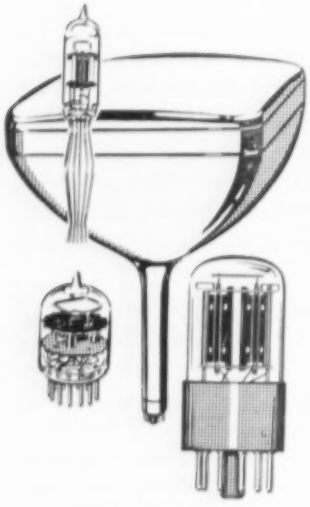
Manufactured in Gt. Britain by

DC 12 PRESSURE UNIT (WITH TRANSFORMER)
As DC 10 but fitted with totally enclosed 100v line transformer tapped 2·5, 5, 8 and 12 watts.
PRICE £7 10s. 0d.

Rola Celestion Ltd.
FERRY WORKS THAMES DITTON, SURREY
Telephone: EMBerbrook 3402

If you manufacture ELECTRON TUBES

LUMA supplies material



Manufacturing all types of incandescent, mercury vapour, neon and flourescent lamps, fluorescent fittings and accessories.

90 % of Luma's production of tungsten and molybdenum wire is exported. Luma's wire and coiled filaments for radio valves and other uses are sold in more than 40 countries.

Lumalampan AB runs its own tungsten and molybdenum works in order to have these important elements under its own control. Modern laboratories maintain an efficient production control and ensure progressive technical development.

Consult LUMA about tungsten and molybdenum wire and coiled filaments. Wire is supplied in all dimensions and finishes, e. g. black, cleaned or plated, semi-finished rods or finished electrodes.

Plated wire supplied in the following forms, among others: gold-plated tungsten wire, gold-, silver- or nickel-plated molybdenum wire.

Outstanding properties of LUMA's plated wire:

impermeable coating with good bond to base metal
bright surface
uniform mechanical properties,

which efficiently prevent oxidation and grid emission.

The Luma Factory, the largest of its kind in Scandinavia has high capacity and short delivery times.

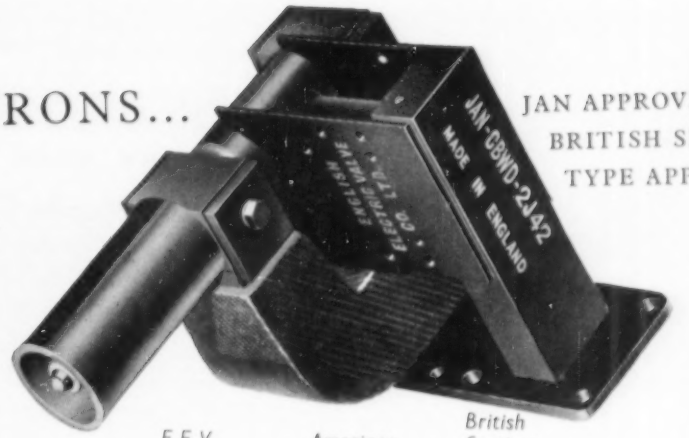
Our tungsten wire catalogue, in English, German or French, will be sent on request.

LUMALAMPAN AB

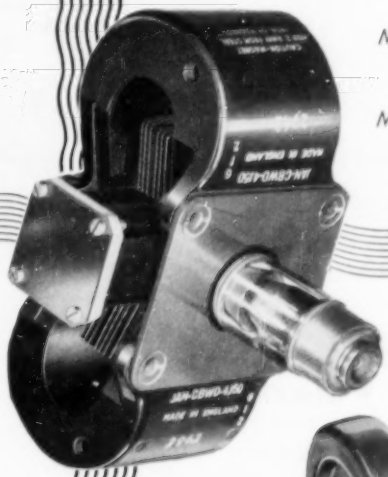
Address: Stockholm 20 Sweden
Cables: LUMALAMPAN Stockholm

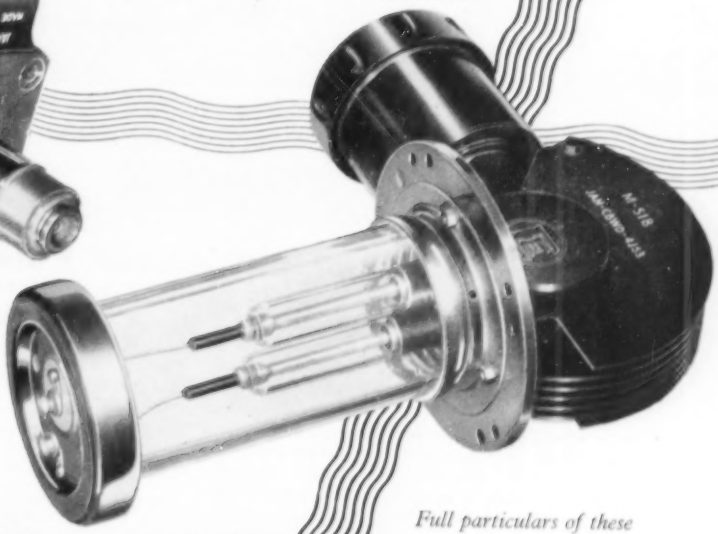
MAGNETRONS...

JAN APPROVED AND
BRITISH SERVICES
TYPE APPROVED



E.E.V. type	American equivalent	British Services type
M.502	4J50	C.V. 2284
M.503		C.V. 1866
	4J31	C.V. 1914
	4J32	
M.518	4J33	C.V. 1916
	4J34	C.V. 1897
	4J35	C.V. 1898
	4J53	C.V. 513
M.526	2J42	C.V. 3676





'ENGLISH ELECTRIC'

Full particulars of these Magnetrons and others in the range available on request.

ENGLISH ELECTRIC VALVE CO. LTD.

Waterhouse Lane, Chelmsford
Telephone: Chelmsford 3491

AP300/41

MUREX SINTERED PERMANENT MAGNETS *are used in this* **ELLIOTT Hermetically SEALED MOVING COIL RELAY**

Where the need for high magnetic stability and efficiency is essential Murex Sintered Permanent Magnets continue to give accurate and reliable service in this and many other applications.

★

MUREX LIMITED
(Powder Metallurgy Division)
RAINHAM · ESSEX
Telephone: Rainham, Essex 3322
Telex 8632. Telegrams: Murex, Rainham-Dagenham, Telex
London Sales Office: CENTRAL HOUSE, UPPER WOBURN PLACE, W.C.1
EUSton 8265

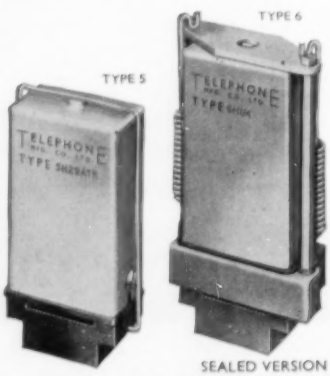
Modern Aircraft...

and the **Carpenter Polarized Relay**

Because rapid acceleration, vibration and wide temperature variations met with in modern aircraft, have little effect on the performance of the smaller versions of the Carpenter Polarized Relay, they are used in many varieties of electronic control and supervisory circuits.

These relays—having a close operate/release differential with operating sensitivities around 0.6 mW responding to frequencies up to 100 c/s—enable the most delicate servo-systems to be operated. Moreover, accurate adjustment is retained for long periods.

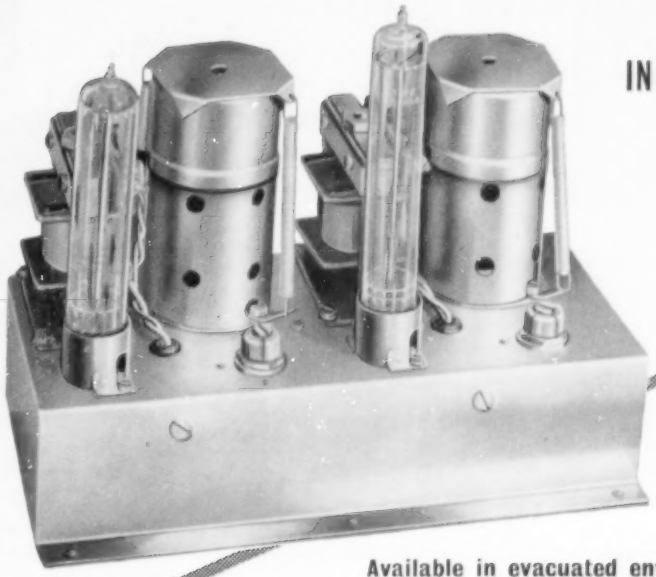
But aeronautics is not the only field in which advantage is taken of the Carpenter Relay's unique qualities—it is being applied very successfully in a wide range of Industrial equipment for *Control, Amplification, Impulse repetition* and *High-speed switching* as explained in our Brochure F.3516—may we send you a copy?



SEALED VERSION

Manufactured by the Sole Licensees:
TELEPHONE MANUFACTURING COMPANY LTD
Contractors to the Government of the British Commonwealth and other Nations
HOLLINGSWORTH WORKS . DULWICH . LONDON S.E.21 TEL. GIPSY HILL 2211

MARCONI LOW FREQUENCY CRYSTALS



IN THE RANGE 1.6-50 Kc/s

The illustration shows a 1.6 and 3 Kc/s Carrier Oscillator Unit for transmitting information to a remote bearing indicator as used in the Marconi V.H.F. Direction Finder Type AD200.

Available in evacuated envelopes with B7G base or with flexible leads for connection into the circuit

Among the many uses for this type of crystal are the following:—

SPEECH INVERSION OSCILLATORS

•

CARRIER CHANNELLING EQUIPMENT

•

RANGE CALIBRATORS

•

TIMING APPLICATIONS ETC.

Flexure mode crystals are suitable for a wide variety of applications where an oscillator of good stability and small size is essential. They cover the frequency range 1.6 to 50 Kc/s with a point of zero temperature coefficient in the range 0°-40°C. The bar in XY flexure used between 1.6 and 12.0 Kc/s has the great advantage of a low temperature coefficient in the region of room temperature.

Lifeline of Communication

MARCONI

MARCONI'S WIRELESS TELEGRAPH COMPANY LIMITED
CHELMSFORD, ESSEX. Telephone: CHELMSFORD 3221

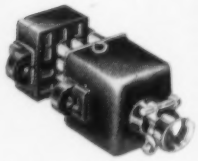
CR4

Better than 5 weeks!

ELCOM

Due to increased production we can now offer
MINIATURE PLUGS and SOCKETS
on short delivery ——

2, 4, 6, 8, 18 & 33 Way ..... ***3 to 4 weeks***
12 & 24 Way ..... ***5 to 6 weeks***



Also

MINIATURE WIRE WOUND VITREOUS & LACQUERED RESISTORS
¾ watt to 25 watt:—
..... ***5 to 6 weeks***

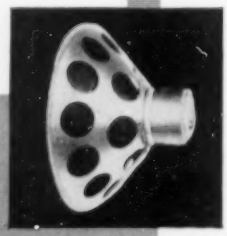
ELECTRONIC COMPONENTS

WEEDON ROAD INDUSTRIAL ESTATE
NORTHAMPTON

TELEPHONE: NORTHAMPTON 2467
GRAMS: "ELCOM" NORTHAMPTON

For a regular
smooth response curve

You need a **PHILIPS**
dual-cone
loudspeaker
(Made in Holland)



A special dual-cone design distinguishes Philips high fidelity speakers, resulting in energy transmissions that are almost independent of frequency. This ensures that, in an ordinary room, sound pressure within an angle of 90 degrees does not vary by more than six decibels; while the excellent spatial distribution of acoustic energy — even at the highest frequencies — is obtained by Philips choice of coupling factor between high-range and low-range cones.

The small cone acts as a high note radiator for frequencies *above 10,000 cycles* and also as a diffuser for frequencies below 10,000 cycles from the large cone. The large cone itself acts as a low note radiator *below 10,000 cycles*, and as a reflector for the high notes above this frequency. The distribution of sound over the entire frequency range is thus much wider than on a normal loudspeaker. *These loudspeakers have a very smooth response curve combined with a low resonance frequency.*

The Philips dual-cone loudspeaker comes in two sizes: 8" and 12", price 6½ gns. (*tax paid*) and 10 gns. *respectively. There is also a single-cone loudspeaker, available in the same two sizes:* price £6.2s.6d. (*tax paid*) and £10.0s.0d. *respectively.*

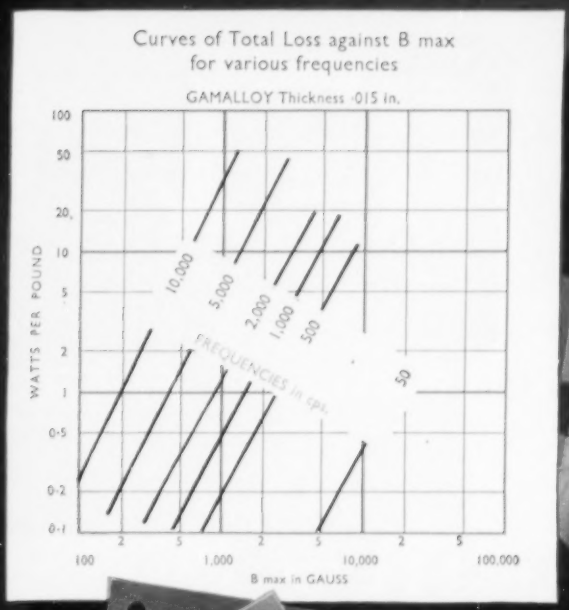
N.B. Any of these speakers may be used on their own or with another suitable loudspeaker using a crossover unit.

For full details write to:

PHILIPS

PHILIPS ELECTRICAL LTD
Musical Equipment Department · Century House · Shaftesbury Avenue · London · W.C.2.

(PR437)



GAMALLOY: nickel-iron core material

'GAMALLOY 48' is the newest nickel-iron core material to be added to the wide range of R.T.B. magnetic materials. As the inset curves indicate, GAMALLOY has low total losses throughout the audio frequency range and, moreover, its permeability at flux densities up to 14,000 gauss is very good. In every way GAMALLOY is an ideal core material for use in small transformers and chokes required in the telecommunications and electronics industry.

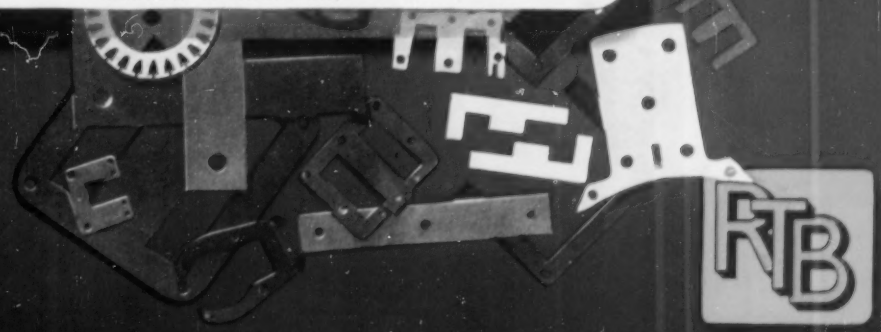
GAMALLOY is manufactured as hot-rolled sheets in widths up to 17 inches, and in two standard thicknesses—·015" and ·005".

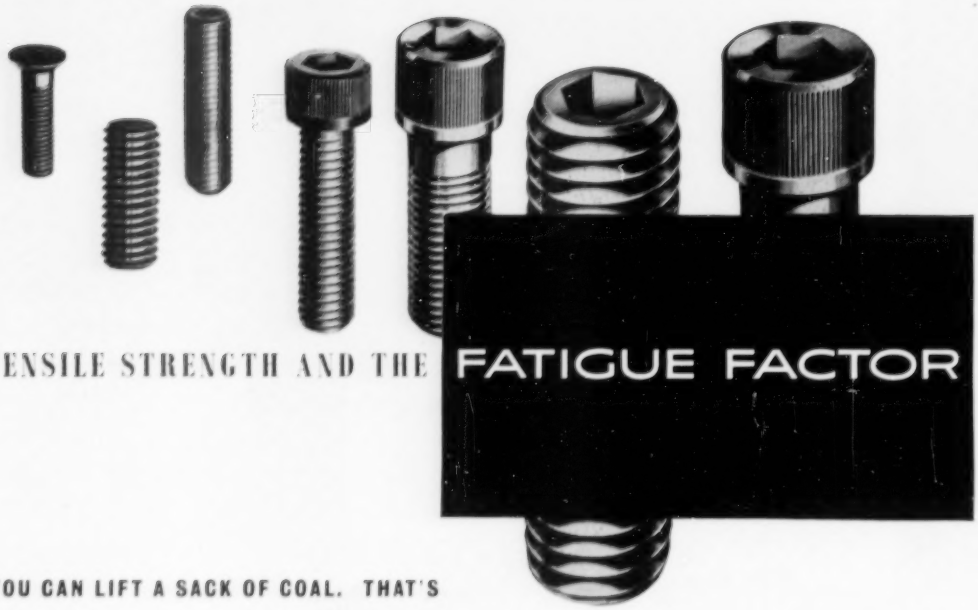
Full technical data can be found in our new catalogue.

RICHARD THOMAS & BALDWINS LTD.

LAMINATION WORKS: COOKLEY WORKS, BRIERLEY HILL, STAFFS
MIDLAND SECTION OFFICE: WILDEN, STOURPORT-ON-SEVERN, WORCS
HEAD OFFICE: 47 PARK STREET, LONDON, W.1

Our Cookley Works is one of the largest in Europe specializing in the manufacture of laminations for the electrical industry.



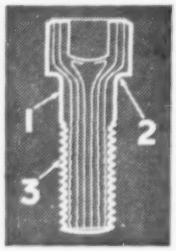


TENSILE STRENGTH AND THE FATIGUE FACTOR

MAYBE YOU CAN LIFT A SACK OF COAL. THAT'S STRENGTH, BUT WHEN YOU'VE BEEN DOING IT ALL DAY, THAT'S FATIGUE.

It has always been an easy matter to show a high Rockwell reading of hardness by merely heat treating any one of many varied steels. These steels may show up splendidly in a tensile test, but are completely worthless on today's higher speed machines and heavier stress forming service conditions. Why are they worthless? They have no fatigue life. Fatigue is a factor which has only received full consideration in recent years. It simply means that metal gets tired and gives way under continual stress, as when you bend a piece of wire back and forth until it breaks.

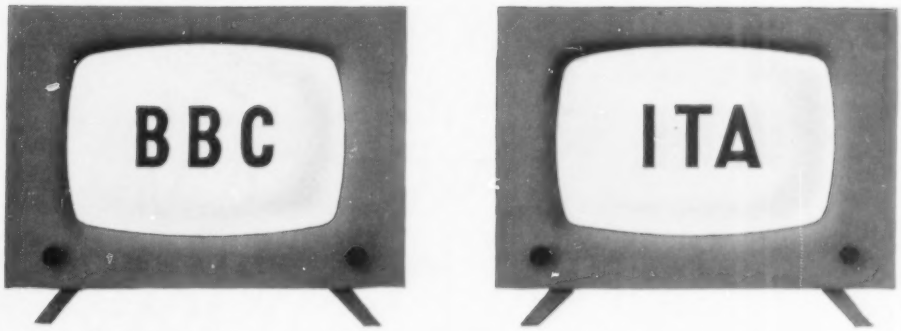
Unbrako technicians constantly strive to achieve the perfect balance between tensile strength and fatigue resistance. That is why Unbrako use only best quality Alloy Steel to their own exacting specifications, and make screws to the highest possible standards of uncarying precision. At the trouble spot under the head (1) Unbrako have a fatigue-resisting fillet. At the thread root (3), fatigue resistance is greatly increased by the continuous closely-knit grain flow and rolled thread (2) and superior notch-free surface finish. No doubt, Unbrako make the world's finest screws and you can specify them with confidence. May we send you fuller details?



UNBRAKO

UNBRAKO SCREWS COST LESS THAN TROUBLE

THE UNBRAKO SOCKET SCREW COMPANY LIMITED · COVENTRY



Two Programmes on one cable

With the minimum of complexity

Our manufacturing specification for television relay cables ensures very low cross-talk couplings within the cable. This provides for satisfactory common carrier operation with consequent simplification of system design. Further details of the BICC standard range of TR8 cables are contained in Publication TD T20 available on request.

BICC

TELEVISION RELAY CABLES

Polythene insulated and sheathed, screened quad cables.

BRITISH INSULATED CALLENDER'S CABLES LIMITED, 21 BLOOMSBURY STREET, LONDON, W.C.1

Consistency of Performance

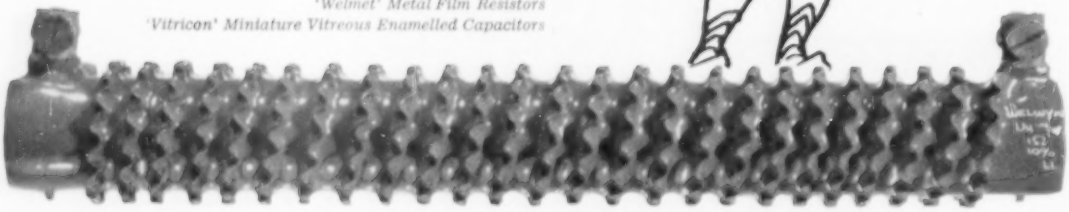
OVERLOADED BUT UNDISMAYED

Before they leave the factory, batch samples of Welwyn heavy duty resistors are put through their paces in no uncertain manner. Nine times the normal load is passed through. But even this abuse does not daunt the Welwyn heavy duty resistor. It is so designed to involve the maximum, practical weight of resistance element. This means a very high thermal capacity which in turn permits absorption of considerable overcharge surges. The winding tape is protected from the possibility of shorted turns or hot spots by a vitreous enamel coating. Other important features include: a very high intermittent rating, availability in adjustable form and in resistance values as low as 0.1Ω.

MANUFACTURERS OF
WELWYN
ELECTRICAL
COMPONENTS

WELWYN PRODUCTS INCLUDE:

Toroidally Wound Power Potentiometers
Panclimatic High Stability Carbon Resistors
Vitreous Enamelled Wirewound Resistors
Carbon Composition and Wirewound Potentiometers
Insulated High Stability Carbon Resistors
High Voltage Composition Resistors
'Welmet' Metal Film Resistors
'Vitricon' Miniature Vitreous Enamelled Capacitors



WELWYN ELECTRICAL LABORATORIES LIMITED · BEDLINGTON · NORTHUMBERLAND

On Admiralty, Ministry of Supply (A.I.D. Approved), and Post Office Lists

High Slope Beam Tetrode

for wide-band amplifiers

Characteristics

High Figure of Merit

TYPE : 5A/170K

Heater Voltage :	6.3	V
Nominal Current :	0.3	A
Mutual Conductance :	16.5	mA/V
Input Capacitance :	7.9 ± 0.6	pF
Output Capacitance :	2.9 ± 0.4	pF
Max. Direct Anode Voltage :	210	V
Max. Direct Anode Dissipation :	3.3	W
Max. Direct Screen Voltage :	175	V
Max. Direct Screen Dissipation :	0.9	W
Max. Direct Cathode Current :	25	mA

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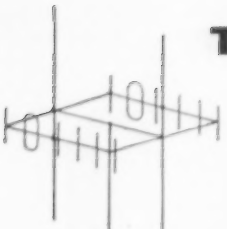

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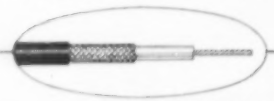

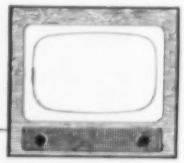

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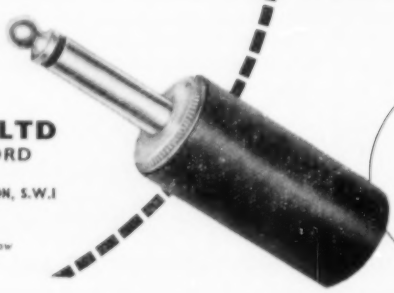

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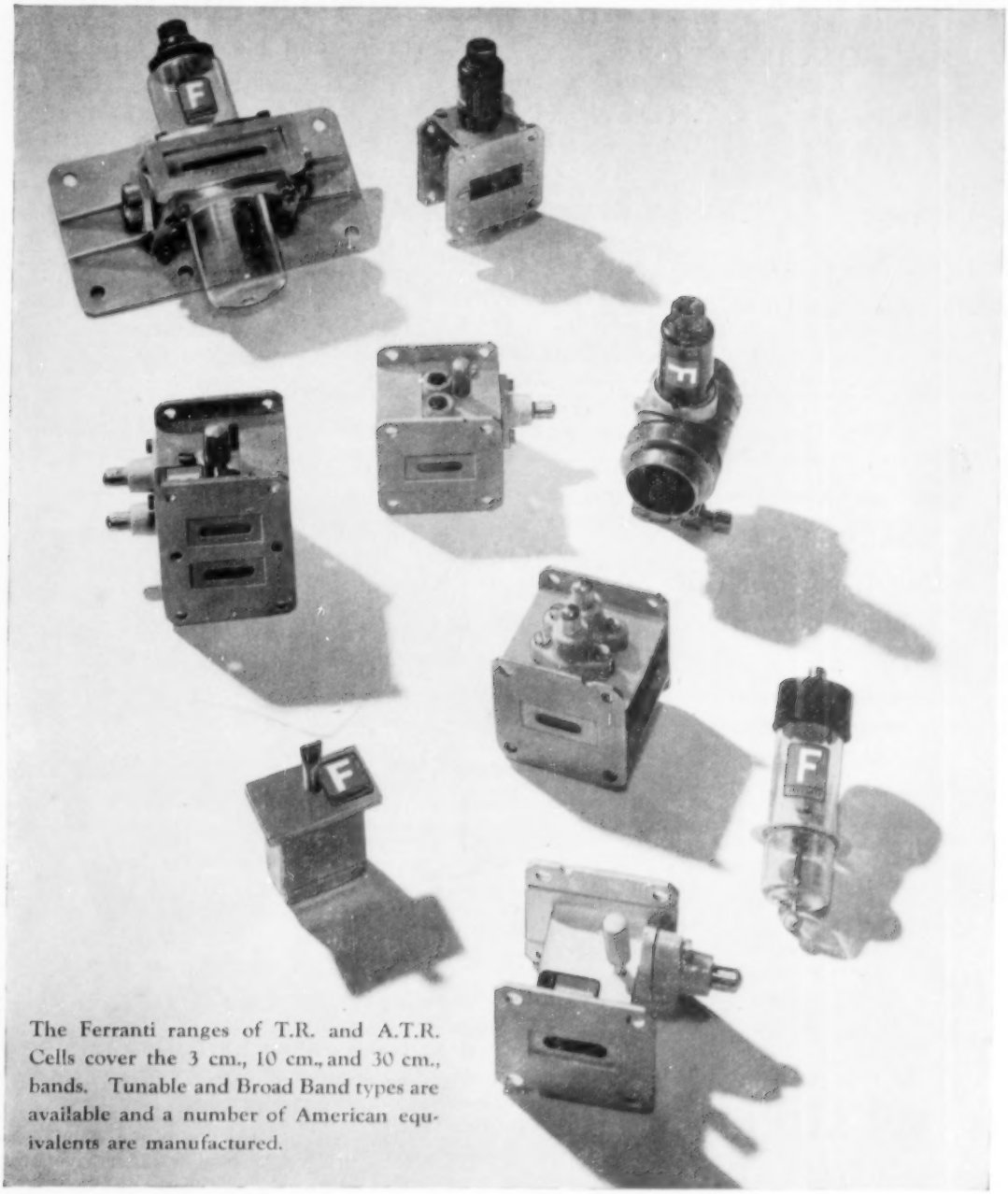

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SPECTRUM OF TELEVISION SIGNALS

By D. A. Bell*, Ph.D., M.I.E.E. and G. E. D. Swann†, M.Sc.

(* University of Birmingham. † Formerly University of Birmingham, now with Marconi's Wireless Telegraph Co. Ltd.)

1. Introduction

THE distribution of power over the sidebands of a television signal is of practical importance in at least three engineering problems; namely, the allocation of frequency channels to television transmitters, the development of 'compatible' colour-television systems, and the use of channels near or in the television band for other services in different geographical regions. In relation to the last, it should be pointed out that very low field-strength can be used for narrow-band communication, so that such a communication channel might suffer interference from a high-power television transmitter situated at a considerable distance, without itself being capable of causing interference in the service area of the television transmitter. Moreover, the aerials for a v.h.f. communication system will not normally be as high as those of television broadcasting stations, and this also will tend to prevent the communication signal from reaching a (low) receiving aerial in the television area but not vice versa. Following a theoretical review[1] and some sample measurements on the low-frequency sidebands[2] by one of the authors, more systematic measurements have been made on the signals broadcast from the Sutton Coldfield television station.

2. Experimental Apparatus

At the receiving site in Birmingham the signal/noise ratio was good, and a t.r.f. receiver was used with a tuned stub in the aerial circuit as rejector filter for the sound transmission. The overall response of the receiver is shown in Fig. 1: this response curve was available to correct readings of the output meter to true magnitude of signal component, but was hardly necessary since the response was within $\pm$ 1·5 dB over the range of 0–2·5 Mc/s. The standard condition adopted was a rectified output from the receiver corresponding to 0·8 V on peak white, when receiving a vision signal having a total excursion of 70% of peak-white carrier amplitude: this implied 560 mV video excursion.

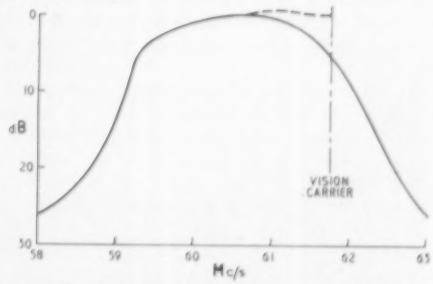

Fig. 1. *Pass-band of r.f. equipment. The full line gives the measured response and the dotted line the equivalent video response (vestigial-sideband transmission).*

In the part of the video-frequency band from zero to 11 kc/s the spectrum was analysed with a wave-analyser having a bandwidth of 4 c/s, thus making it possible to resolve the various multiples of the 25-c/s picture frequency. The difficulty

MS accepted by the Editor, November 1955

remained that it was impracticable to identify individually the 25-c/s components in the kc/s range, particularly as the B.B.C. scanning system was related to 'grid' frequency so that the individual harmonics were not precisely fixed in frequency. This difficulty was overcome by fitting a slow-speed motor drive to the tuning mechanism of the wave-analyser and connecting its output to a recording meter. A 'comb' spectrum of the harmonics then appeared on the recorder chart, and individual harmonics could be identified by counting through the series either from zero frequency or from the very strong component at the line frequency (10·125 kc/s nominal). Since the component amplitudes are a function of picture content, which is a more or less randomly varying function, 20 such records were taken as a basis for averaging.

As it was not practicable to extend this process over the whole range of the signal, comprising some 10^5 harmonics of 25 c/s, the general trend of the spectrum was studied by exploring it with a communication receiver in the ranges 90 to 550 kc/s and 1·38 to 1·97 Mc/s. The receiver was set to its narrowest i.f. band, the selectivity characteristic of which is shown in Fig. 2, and this was tuned in turn to each harmonic of the 10·125-kc/s line frequency. (It follows from Fig. 2 that the receiver output would include partially the effects of a dozen or so of the components which, in practice, are spaced on either side of the line harmonic by multiples of the frame frequency.) A variable attenuator was inserted between the video output and the input to the communication receiver so that, at each frequency, the detector diode current of the latter could be set to a standard value and the attenuation compared with that required with an input of 1 mV from a signal generator. An 'average' value for each component was obtained by watching the output meter for a short period of time and estimating a mean value.

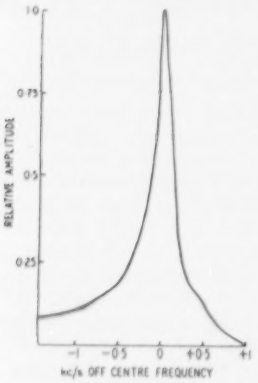

Fig. 2. *Response of narrow-band receiver used for analysis of spectrum.*

3. Analysis of Observations

The fine structure of the first 11 kc/s of the video spectrum is illustrated in Fig. 3*. The intervals between the lines are of course not completely devoid of signal, but earlier experiments[2] suggest that the residual signal is of the order of 30 dB below the lines. Since Fig. 3 represents average values, it is important to examine also the dispersion of the individual readings within the average. In the hope that the distribution might be Gaussian, the individual observations for each of several components were plotted on arithmetic probability paper[3] as shown in Fig. 4. The components at 50 c/s and 1,100 c/s

* Reproduced from "Information Theory and its Engineering Applications", 2nd Edition, Pitman, 1956.

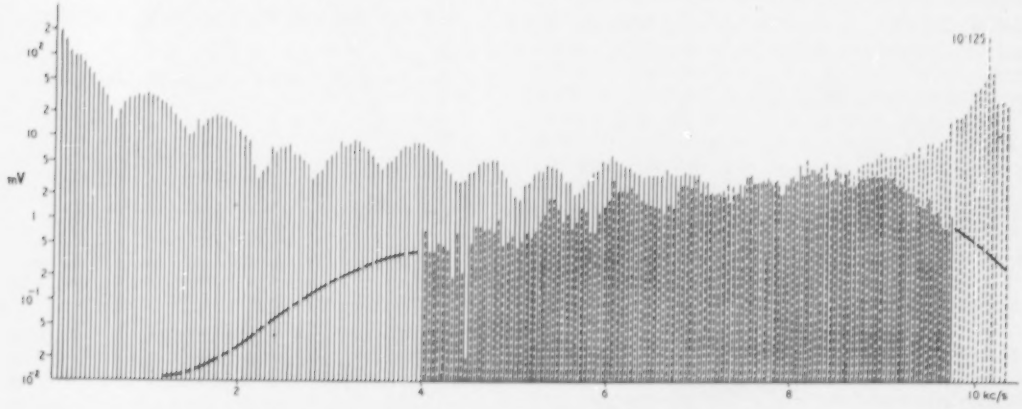

Fig. 3. *Fine structure of video spectrum in range 0–10·2 kc/s.*

in Fig. 4(a) can be fitted to a single straight line as well as to anything; but, in the light of the very definite shape of distribution at 9,175 c/s in 4(c), the intermediate frequencies in 4(b) have been given three-slope distributions.

For interference predictions, it is important to know the probability of occasional large values, as well as the mean values. In the absence of a simple distribution law, rough estimates have been obtained by extrapolating the distribution graphs of Fig. 4 as though they were Gaussian in the outer parts; Table 1 shows the ratio

(i) The amplitudes decrease faster than $1/n$, the products for the highest harmonics being about $\frac{1}{4}$ of the value that would be predicted from the lower components and the $1/n$ rule.

(ii) If the wave under analysis were a simple rectangular wave (of the type described in a previous paper[1] as a 'line-frequency sub-carrier') every tenth harmonic would be missing. Since these tenths were not of appreciably different amplitudes from other harmonics, the spectrum must be dominated by the picture information and not by the synchronizing pulses.

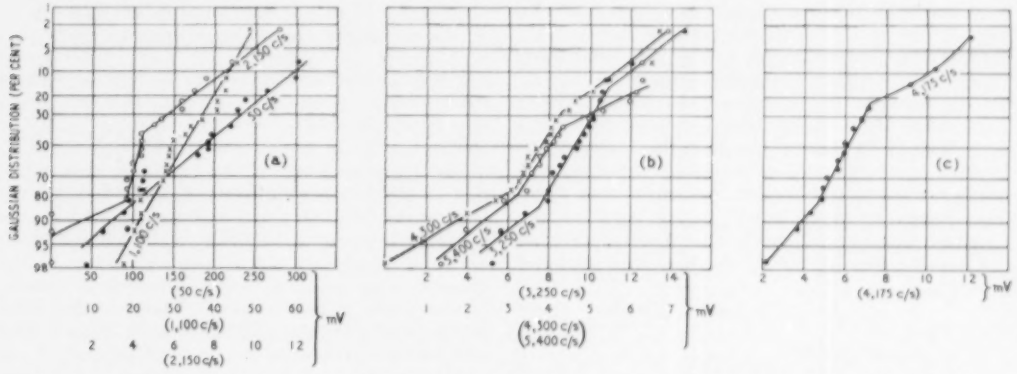

Fig. 4. *These curves show individual observations for specific frequencies plotted on arithmetic probability paper.*

to the mean of a 'peak' amplitude defined as that amplitude which would then have 0·1% probability.

TABLE 1

Frequency (c/s)	50	1,100	2,150	3,250	4,300	5,400	9,175
Peak/mean	2·5	1·8	3·7	2	2·3	2·9	2·7

It should be noted that the law is uncertain and the extrapolation is rather large—effectively from $2\frac{1}{2}$% to 0·1% probability—and these figures should be taken to indicate only an order of magnitude of peak-to-mean of say 3 : 1.

When a waveform is subjected to a Fourier-series analysis, the amplitudes of the higher harmonics are expected to show a general trend of diminution at least as fast as the order of the harmonic. In order to check the conformity of the television spectrum with the rule, the amplitudes of line-harmonic components were measured as described in Section 2 for the ranges 9 to 54 and 136 to 195 times the line frequencies. The amplitudes of the harmonics were then multiplied by their individual order numbers so that for an exact $1/n$ law of amplitudes the result would have been constant, and the results are plotted in Fig. 5. The following are the notable features of this diagram.

4. Relation between Peak-White Power and Sideband Power

From the point of view of interference with other services, it is important to be able to estimate for a given power of television transmitter the power in any designated section of the video spectrum; i.e., the power that would be radiated into a narrower frequency band which was within the sideband range of the television spectrum.

Since the picture waveform is unpredictable there is no specific answer to this question. But a reasonable limit, which is not likely to be exceeded too often, is set by first assuming a full white raster, so that the transmitter radiates peak-white power except during synchronizing pulses, and then taking the envelope of the $(\sin x)/x$ distribution which arises from the Fourier analysis of a rectangular wave. Frame sync pulses will be ignored*, and the waveform sketched in Fig. 6 can be represented by

$$s = (1 - x)\,a + \frac{2a}{\pi} \sum_{m} \frac{1}{m} \left[\sin 2\pi m \left(\frac{1 - x}{2} \right) \right] \cos 2\pi m l t \qquad \ldots\ldots (1)$$

* See Reference 4 for a method of obtaining the exact spectrum.

where l is the line frequency and m the order of harmonic. On the second assumption, taking the envelope of $(\sin x)/x$, the sideband part of the series reduces to

$$\frac{2a}{\pi}\sum_{m=1}^{n}\frac{1}{m}\cos 2\pi mlt \quad .. \quad .. \quad .. \quad (2)$$

the series ending after n terms where n is the number of line harmonics in the whole video band. It is, of course, impossible for all components to appear with these amplitudes simultaneously, but the detailed structure of the picture at any moment will enhance certain harmonics at the expense of others, and the expression (2) is an estimate of the amplitudes which may occur from time to time with reasonable pictures. It would not be relevant to the transmission of a picture consisting of a single homogeneous resolution test, such as the classic but hypothetical chequer board; however, neither practical resolution test cards nor programme pictures are of this type. The power is proportional to the sum of the squared amplitudes, with the mean-square of $\cos 2\pi mlt$ replaced by $\frac{1}{2}$, so that the sideband power is

$$P_s = \frac{2a^2}{\pi^2}\sum_{m=1}^{n}\frac{1}{m^2} \quad .. \quad .. \quad .. \quad (3)$$

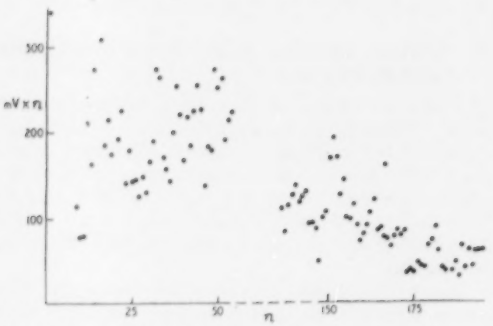

Fig. 5. *Plots of the product of harmonic amplitude and order of harmonic.*

When n is 200 the error resulting from taking the sum to infinity instead of the sum to n terms is approximately 1/200 so that, for practical purposes, it suffices to take the sum to infinity, which is $\pi^2/6$. It follows that $P_s = a^2/3$ when the peak-white power is a^2. Note that, whereas a double-sideband amplitude-modulated wave analyses into components which individually have an amplitude factor of $\frac{1}{2}$, the vestigial-sideband system doubles the amplitudes of all but the lowest components in the retained sideband, so that the amplitude of the component is equal to the amplitude of the corresponding harmonic in the analysis of the modulating waveform.

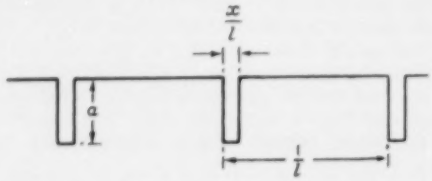

Fig. 6. *Rectangular waveform.*

When a limited channel in the upper part of the spectrum is to be considered, it should be a fair approximation to multiply the power in the central harmonic within the band by the number of line harmonics within the band. For example, in a channel extending from 1·6 to 1·75 Mc/s from the carrier frequency of a B.B.C. transmitter, the central harmonic of 10·125 kc/s will be the 165th and the number of harmonics will be 14 if the channel is very closely defined or, more probably, 15 with the rate of cut-off likely in practical receiving apparatus. The power in this channel will then be estimated as

$$15 \times \frac{2}{\pi^2} \times \frac{1}{165^2} \times \text{peak-white power}$$

or about 40 dB below peak-white power.

On the basis of equation (1) and the argument used above, the maximum ordinate to be expected in Fig. 6, based on 800 mV, for peak white, would be 510. It appears that, in practice, this argument gives a factor of safety of 6 to 10 dB over *typical* values of the sideband power but, since instantaneous values may be 10 dB above the mean value (cf. Table 1, for fine-structure components, in Section 3) the result obtained seems to provide a reasonable basis for design.

5. Acknowledgments

The experimental work was carried out in the Electrical Engineering Department, University of Birmingham, and the authors wish to acknowledge the facilities placed at their disposal by Professor A. Tustin, some time head of the Department.

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ATTENUATORS WITH MISMATCHED TERMINATIONS

By J. Altshuler, E.E.(Havana)

THE loss of power in a four-terminal network is known as attenuation and is often expressed in decibels. If the four-terminal network is wholly made up of resistive elements, it is called an attenuator and this, in turn, is said to operate under nominal conditions when used to couple two resistance levels which match its image resistances. A formula for the attenuation introduced by an attenuator under non-nominal conditions will be developed here.

To begin with, consider a resistive, dissymmetrical attenuator, defined by (a) its image resistances, R_1, R_2, and (b) its image attenuation constant, α, such that

$$e^{\alpha} = \sqrt{\frac{E_{1i}I_{1i}}{E_{2i}I_{2i}}} \qquad .. \quad .. \quad .. \quad (1a)$$

where, with reference to Fig. 1, E_{1i} and E_{2i} represent, respectively, the input and output r.m.s. voltages to the attenuator. I_{1i} and I_{2i} are the corresponding r.m.s. currents, while the subscript i emphasizes the fact that all the voltages and currents in (1a) are those produced under image-operation conditions.

If we make

$$N = e^{\alpha} \qquad .. \quad .. \quad .. \quad .. \quad (1b)$$

then the nominal attenuation in decibels is given by

$$10 \log \left(\frac{E_{1i}I_{1i}}{E_{2i}I_{2i}}\right) = 20 \log N \qquad .. \quad (1c)$$

Since $E_{1i} = I_{1i}R_1$ and $E_{2i} = I_{2i}R_2$, we can write

$$N = \frac{I_{1i}}{I_{2i}}\sqrt{\frac{R_1}{R_2}} \qquad .. \quad .. \quad .. \quad (1d)$$

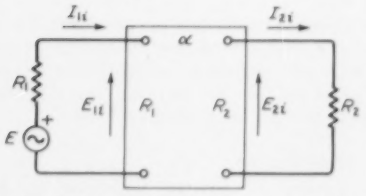

Fig. 1. *Attenuator connected between source and load.*

Input Resistance

Fig. 2 represents the network as defined above connected between a source of internal (resistive) impedance sR_1 and a resistive load of value rR_2, s and r being any two positive numbers whatsoever.

Following the classical treatment by Shea[1] we can substitute parts of the source- and receiving-end impedances by equivalent generators with zero internal impedances and internal e.m.fs of magnitudes and polarities as shown in Fig. 3. This new arrangement does not alter the situation so far as currents and voltages at the terminals of the attenuator are concerned. An important advantage is gained, however, since now image conditions prevail and it is easy to apply the superposition theorem.

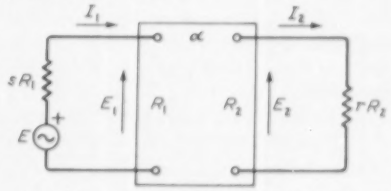

Fig. 2. *Attenuator with mismatched terminating resistances.*

Let I_1' and I_2' be the values which I_1 and I_2 would take if the output generator in Fig. 3 were short-circuited, and let I_1'' and I_2'' be the values which the same currents would take if the input generators were short-circuited.

We have

$$I_1' = \frac{E - I_1(s-1)R_1}{2R_1}$$

and, recalling (1d),

$$I_2' = \frac{E - I_1(s-1)R_1}{2N\sqrt{R_1R_2}}.$$

Also

$$I_2'' = \frac{-I_2(r-1)R_2}{2R_2}$$

and again, from (1d) and the fact that N is independent of the direction of power flow, it follows that

$$I_1'' = \frac{-I_2(r-1)R_2}{2N\sqrt{R_1R_2}}$$

The application of the superposition theorem yields

$$I_1 = I_1' + I_1'' = \frac{E - I_1(s-1)R_1}{2R_1} - \frac{I_2(r-1)R_2}{2N\sqrt{R_1R_2}}$$

$$I_2 = I_2' + I_2'' = \frac{E - I_1(s-1)R_1}{2N\sqrt{R_1R_2}} - \frac{I_2(r-1)R_2}{2R_2}$$

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and solving for I_1 and I_2 we get

$$I_1 = \frac{E}{R_1}\left[\frac{1-m}{s(1-m)+(1+m)}\right] \quad \ldots \quad (2a)$$

where

$$m = \frac{r-1}{N^2(r+1)} \quad \ldots \quad \ldots \quad \ldots \quad (2b)$$

and

$$I_2 = \frac{2\sigma E}{\sqrt{R_1 R_2}\,(s+1)(r+1)N} \quad \ldots \quad (3a)$$

where

$$\sigma = \left[1 - \frac{(s-1)(r-1)}{(s+1)(r+1)N^2}\right]^{-1} \quad \ldots \quad (3b)$$

The input resistance of the attenuator, R_{in}, is obtained immediately since

$$R_{in} = E/I_1 - sR_1$$

and hence, making use of (2a),

$$R_{in} = \left[\frac{1+m}{1-m}\right] R_1 \quad \ldots \quad \ldots \quad (4)$$

m being given by (2b).

Attenuation

Let P_{in} be the power flowing into the attenuator and P_{out} that flowing out of it. By definition

$$\text{Attenuation in decibels} = 10 \log P_{in}/P_{out} \quad (5a)$$

With reference to Fig. 2

$$\text{Attenuation in decibels} = 10 \log \frac{I_1^2 R_{in}}{I_2^2 r R_2} \quad (5b)$$

Now,

$$I_1 = \frac{E}{sR_1 + R_{in}} \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

and substituting this relation and (2b), (3) and (4) into (5) we get, after some algebraic manipulation,

$$\text{Attenuation in dB} = 10 \log \frac{(1+r)^2}{4r} + 10 \log (1-m^2) + 20 \log N \quad \ldots \quad (7)$$

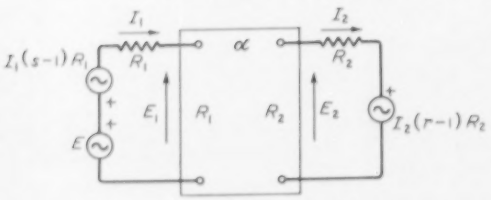

Fig. 3. *Equivalent circuit of Fig. 2.*

Formula (7) shows that the attenuation is independent of the matching conditions at the input terminals of the attenuator. In Fig. 4 the first two terms on the right-hand side of (7) have been plotted against a logarithmic r scale for a number of values of nominal attenuation, as indicated in decibels on the curves. It becomes evident that the relative importance of the second term diminishes very rapidly as 20 log N increases; i.e. for a given value of r and a sufficiently large nominal attenuation, the attenuation in dB is equal to 20 log N plus the reflection loss at the load end

$$10 \log (1+r)^2/4r \quad \text{dB}.$$

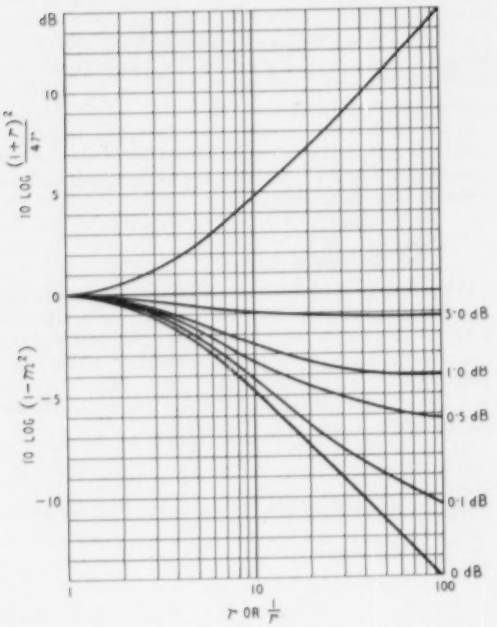

Fig. 4. *Curves of actual attenuation for various amounts of nominal attenuation.*

It is remarkable that (7) is not simplified in the slightest if the attenuator happens to be symmetrical. This should be contrasted with the fact that the often-quoted formula for the insertion loss[1] does contain a symmetry factor.

Acknowledgment

The author is greatly indebted to Mr. M. Estrada for stimulating discussions.

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SCATTERING OF MICROWAVES BY LONG DIELECTRIC CYLINDERS

By Albert W. Adey, Ph.D.

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SUMMARY.—The paper presents near-field results for the scattering of 3-cm waves by dielectric rods of circular, square and rectangular cross-section.

Measured results obtained by the use of a parallel-plate transmission line are in good agreement with calculations of the scattering due to circular polystyrene cylinders having radii comparable with the wavelength. The effect of the material of the cylinder on the scattering is investigated experimentally. The resonances caused by scattering by dielectric cylinders are discussed.

Experimental results are given for square and rectangular polystyrene rods.

Introduction

ALTHOUGH the problem of the scattering of an electromagnetic wave by a circular dielectric cylinder was studied many years ago by Rayleigh[1], the present author knew of only one previous attempt[2,3] to compare calculated with experimental results under reasonably comparable conditions when the work discussed in this paper was begun. The study referred to above considered the case of damped waves scattered by thin cylinders. More recently, other calculated and experimental results have been announced[4]. The recent development of the parallel-plate transmission line has permitted better controlled microwave experiments with cylindrical obstacles, for some cross-sections of which theoretical scattering solutions are not yet available.

In a recent paper[4] where he presented results for metal cylinders, the author described such a transmission line operating at a wavelength of 3·275 cm. In the present paper corresponding results for dielectric rods of circular, square and rectangular cross-section will be given.

In contrast to scattering by metal cylinders, it is shown that resonances can occur when the scattering is by dielectric cylinders, because of the field's penetration of the material. These effects are illustrated by calculated and experimental results. Measurements of the scattered fields of cylinders of various dielectric materials show the damping effect resulting from the loss in the dielectric.

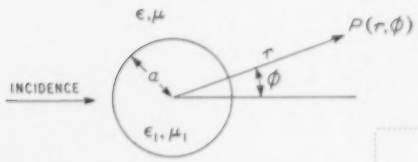

Fig. 1. *Section through the cylindrical co-ordinate system.*

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Scattering by a Circular Cylinder

A cylinder of radius a permittivity ϵ_1 and permeability μ_1 is illuminated by a plane wave, of amplitude unity for convenience, which is travelling in the direction $\phi = 0$ and which has its electric vector polarized parallel to the cylinder axis. Fig. 1 shows the arrangement, for a cylindrical system of co-ordinates (r,ϕ,z). The medium external to the cylinder is considered to be characterized by a permittivity ϵ and permeability μ. A time factor $e^{-j\omega t}$ is assumed and rationalized m.k.s. units are used.

Following a standard procedure, the incident, scattered and internal fields may be written in the respective forms—

$$E_{inc}(r,\phi) = e^{jkr} = \sum_0^\infty e_n (j)^n \, \mathrm{J}_n(kr) \cos n\phi \quad (1)$$

$$E_{sc}(r,\phi) = \sum_0^\infty e_n \, B_n \, \mathrm{H}_n^{(1)}(kr) \cos n\phi \quad (2)$$

$$E_{int}(r,\phi) = \sum_0^\infty e_n \, A_n \, \mathrm{J}_n(k_1 r) \cos n\phi \quad (3)$$

where the J_n are the Bessel functions of the first kind, the $\mathrm{H}_n^{(1)}$ are the Hankel functions of the first kind and e_n is Neumann's number ($e_o = 1$; $e_n = 2$, $n = 1, 2 \ldots$)

$k = 2\pi/\text{external wavelength}$

$k_1 = 2\pi/\text{internal wavelength}$

$$= \sqrt{\frac{\mu_1 \epsilon_1}{\mu \epsilon}}\, k = k_r k$$

From Maxwell's equation for the magnetic field H_ϕ —

$$H_\phi(r,\phi) = \frac{j}{\omega\mu} \frac{\partial}{\partial r} E(r,\phi) \quad .. \quad .. \quad (4)$$

and from the continuity of the tangential electric and magnetic fields at the boundary $r = a$, the amplitude of the scattered field can be obtained as

$$B_n = -(j)^n \frac{\frac{k}{\mu} \mathrm{J}_n(k_1 a)\, \mathrm{J}_n'(ka) - \frac{k_1}{\mu_1} \mathrm{J}_n(ka)\, \mathrm{J}_n'(k_1 a)}{\frac{k}{\mu} \mathrm{J}_n(k_1 a)\, \mathrm{H}_n^{(1)\prime}(ka) - \frac{k_1}{\mu_1} \mathrm{H}_n^{(1)}(ka)\, \mathrm{J}_n'(k_1 a)} \quad (5)$$

where the prime indicates differentiation with respect to the argument. When $\mu = \mu_1$ this can be reduced to—

$$B_n = -(j)^n \frac{C_n}{C_n + jD_n} \quad .. \quad .. \quad .. \quad (6)$$

where

$$\left.\begin{aligned} C_n &= \mathrm{J}_n(k_1a)\ \mathrm{J}_{n+1}(ka) - k_r \mathrm{J}_{n+1}(k_1a)\ \mathrm{J}_n(ka) \\ D_n &= \mathrm{J}_n(k_1a)\ \mathrm{Y}_{n+1}(ka) - k_r \mathrm{J}_{n+1}(k_1a)\ \mathrm{Y}_n(ka) \\ k_r &= \sqrt{\epsilon_1/\epsilon} \end{aligned}\right\} \quad (7)$$

Similarly, the coefficient A_n of Equ. (3) is found to be—

$$A_n = (j)^n \frac{k}{\mu} \frac{[\mathrm{J}_n(ka)\ \mathrm{H}_n^{(1)\prime}(ka) - \mathrm{J}_n'(ka)\ \mathrm{H}_n^{(1)}(ka)]}{\frac{k}{\mu}\mathrm{J}_n(k_1a)\ \mathrm{H}_n^{(1)\prime}(ka) - \frac{k_1}{\mu_1}\mathrm{J}_n'(k_1a)\ \mathrm{H}_n^{(1)}(ka)} \quad .. \quad .. \quad (8)$$

which reduces, for $\mu = \mu_1$, to—

$$A_n = (j)^{n-1} \left(\frac{2}{\pi ka}\right) \Big/ \left(C_n + jD_n\right) \quad .. \quad (9)$$

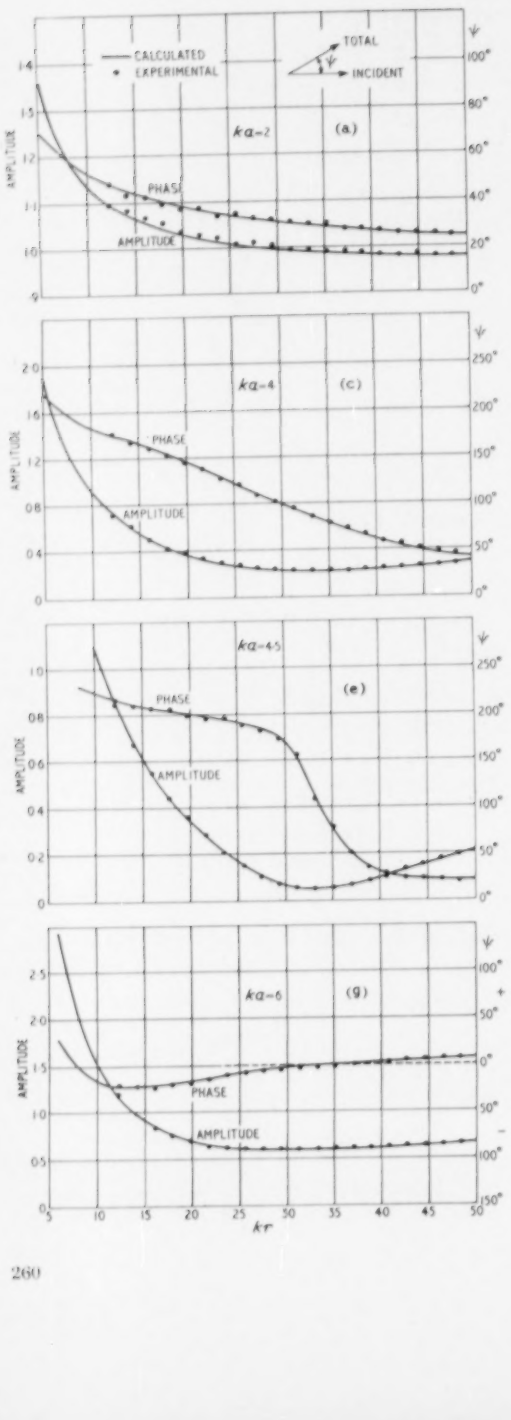

Fig. 2. *Comparison of measured and calculated shadow fields for a series of values of ka.*

The total external field is then given as the sum of the incident and scattered fields by Equs. (1), (2) and (5) or (6).

(a) Scattering by a Circular, Polystyrene Cylinder

Calculations have been made of the near forward-scattered ($\phi = 0$) and back-scattered ($\phi = \pi$) fields for several values of ka, for $\mu = \mu_1$ and $k_r = 1{\cdot}60$. The total shadow fields are compared with experimental results in Fig. 2.

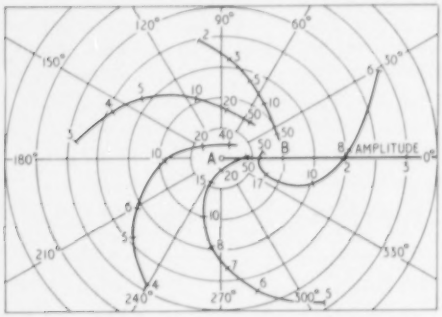

Fig. 3. *Vector diagram illustrating the trend of the curves of Fig. 2. The numbers on the curves refer to values of kr, those at their ends to values of ka.*

The reference for amplitude is the incident field (essentially plane). The plotted phase for a particular value of kr is the shift in phase at the point corresponding to that value of kr on introducing the cylinder into the field.

Three points of difference between the shadow fields presented here and those given previously for metal cylinders are immediately evident—

(1) While the field amplitude approached zero as the field point approached the surface of the metal cylinder, it showed a sharp rise above the incident field value near the polystyrene cylinder surface.

(2) While the amplitude increased monotonically up to the incident field value for increasing values of kr in the metal cylinder case, for the polystyrene cylinders it first decreased before finally increasing to the incident field value.

(3) For the ka values above about 4·6 there appears to be a discontinuity of 2π in the phase for the larger kr values.

The trends indicated by these curves can best be illustrated by combining, for each cylinder, the amplitude and phase into a vector diagram. This has been done for several values of ka in Fig. 3. The numbers on each curve indicate the kr value of the field points, the number at the outer

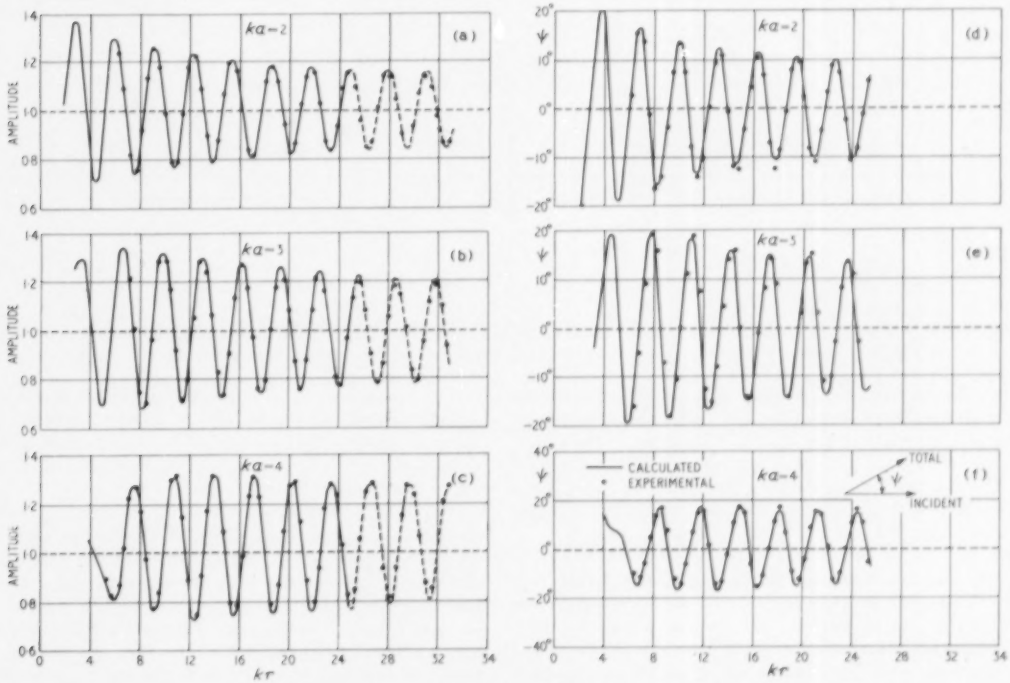

Fig. 4. *Field on the illuminated side of cylinders.*

extremity of each curve indicating the *ka* value of the cylinder. The shadow deepens until that value of *ka* is reached for which the curve passes through the origin, after which it decreases. For that value of *kr* and *ka* the amplitude is zero and the phase indeterminate. The zero-field region is very localized with respect to *ka* and *kr*. For *ka* values less than the critical the phase decreases with increasing *kr*. For greater *ka* values the phase first decreases and then increases, tending toward 2π instead of zero. This behaviour was suspected by Kodis[5] in his experiments on dielectric cylinders, although he doubted whether it would occur on the shadow axis.

corresponding calculations were not done, however, mainly because of the lack of tables of Bessel Functions with complex argument. The results have been presented in polar form, rather than in field-point detail, in Fig. 5 for the following materials respectively—polyethylene, graphite, and two resin-base plastics.

Because of the lower value of relative permittivity of polyethylene, the curves of Fig. 5 (a) lie at lower angles than the corresponding ones of Fig. 3 for polystyrene. A higher value of *ka* is thus required of the polyethylene cylinder to give the zero-field phenomenon than was found for polystyrene.

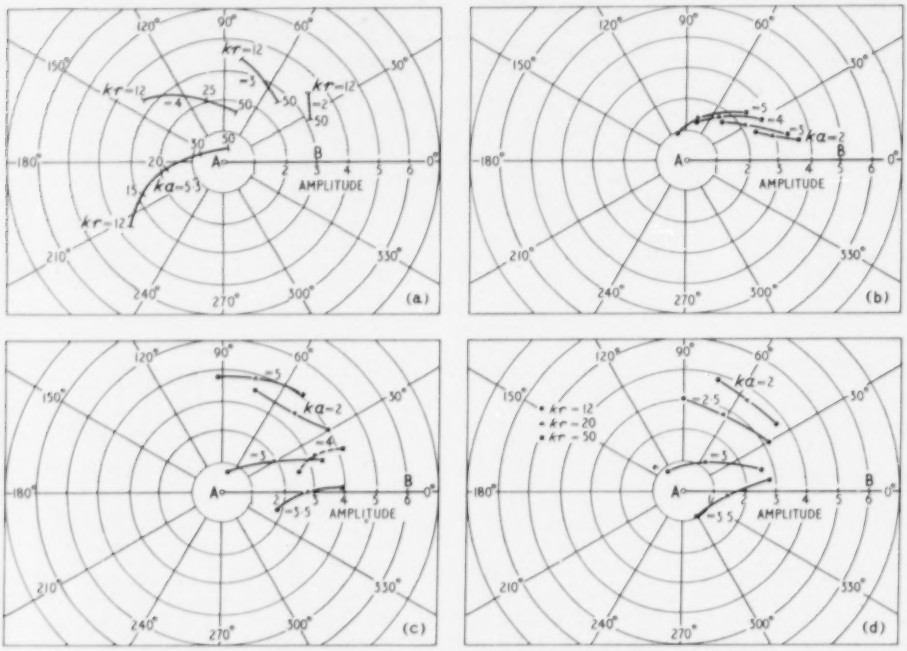

Fig. 5. *Shadow fields for cylinders of polyethylene (a), graphite (b) and resin-based plastics (c and d).*

The curves of Fig. 4 for the field on the illuminated side of the cylinders follow, in the main, the pattern for the metal cylinders. However, the amplitude of the standing wave does not always fall off monotonically with increasing *kr* as was found in the metal case, but can actually rise, initially.

(*b*) *Measurements on Miscellaneous Materials*

The shadow measurements were repeated for families of cylinders of several other readily-available materials, to determine the effect on the scattering of the change of permittivity together with the presence of appreciable loss. The

The results for graphite cylinders show that the graphite and metal cylinders behave similarly for scattering. The scattered field increases with cylinder radius and the left-hand terminal of each curve tends toward the origin.

The effect on the scattering of increasing both the relative permittivity and the loss is illustrated by Figs. 5 (c) and (d). The curve passing through the origin corresponds to a lower value of *ka* than for polystyrene, because of the higher relative permittivity. The higher loss, since it decreases the scattered field amplitude by damping, has the effect of moving the pattern to the right—toward point B.

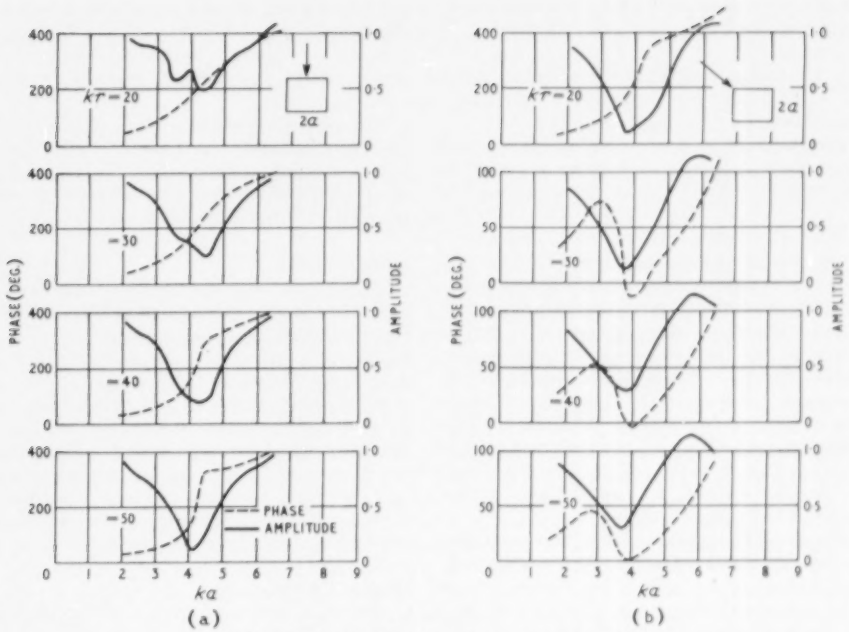

Fig. 6. *Shadow fields for square polystyrene rods.*

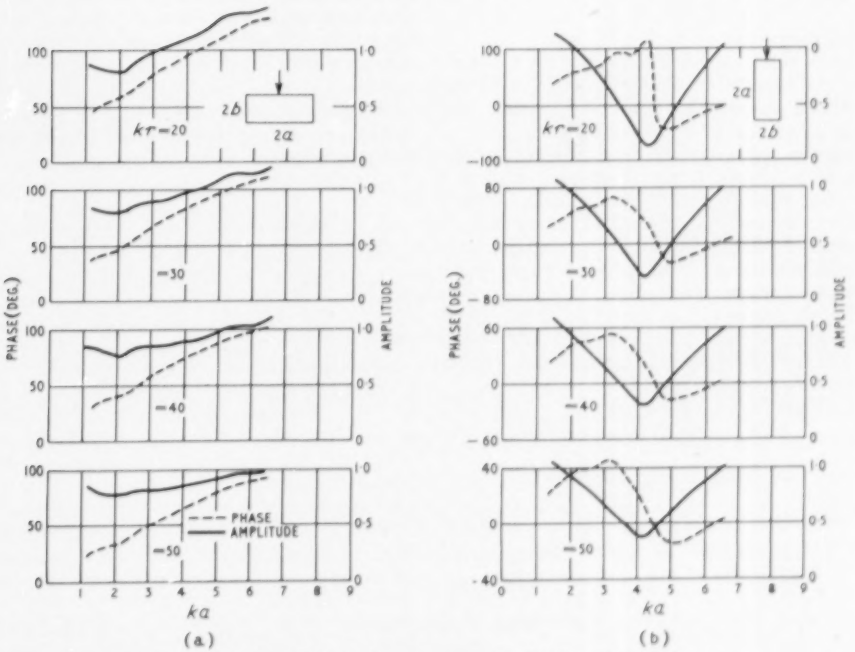

Fig. 7. *Shadow fields for polystyrene rods of rectangular section.*

Scattering by Square and Rectangular Polystyrene Rods

(a) Square Rods

Shadow field measurements were made on a family of ten polystyrene rods in the ka range 2·18–6·43, where "a" is the half-side.

To save space, the results were first plotted in detail as for the circular polystyrene cylinders and then consolidated in Fig. 6 for several values of kr.

For both orientations there is a tendency—observed in the case of the circular cylinders—for the amplitude to have a zero value and the phase a change of form for a particular combination of ka and kr. The field complexity that can be encountered in the near region is well illustrated by the curve for $kr = 20$ for normal incidence.

For normal incidence, the detailed curves indicated that the critical radius-distance combination would be approximately 4·2 and 55 respectively. The ka value is a little smaller than for the circular cylinders, indicating a higher effective relative permittivity.

For diagonal incidence the corresponding values are in the region of 3·7 and 23 respectively.

For convenience the angles on the higher kr curves for diagonal incidence are shown in a different form than for the smaller ones, because of the rapid increase in phase after passing the critical kr value.

(b) Rectangular Rods

Shadow measurements were made on a family of eleven rectangles of dimensions $2a \times 2b$ for $kb = 2{\cdot}37$ and ka in the range 1·52–6·44. The results are given in Fig. 7.

The curves for a constant dimension in the direction of incidence are fairly smooth, although less so than for the metal rectangles treated previously. One would expect the curves of both amplitude and phase to tend, in the limit of large ka, to the values corresponding to the transmission coefficient of a polystyrene slab.

For the other orientation the trend of the curves is again similar to that for the circular and square rods. From the detailed curves the critical combination of ka and kr appeared to be approximately 4·2 and 18 respectively. Thus, as for the squares, a higher value of effective relative permittivity is indicated than for the circular cylinders.

The physical optics aspect of the results (e.g., an analysis based on Fresnel Zone Theory) has not been discussed since, for the largest transverse dimension considered, the field point was always in the first Fresnel Zone.

Conclusions

The study has shown the dielectric, circular cylinders to be resonant structures, their scattering properties varying widely with the cylinder radius and material. This is in contrast with the scattering behaviour of metal rods, which is comparatively smooth as a function of radius. The damping effect of the loss in the cylinder material has been illustrated. The shadow field of the square rods has been shown to be similar to that for the circular cylinders, as has that for the rectangles when the dimension in the direction of incidence is varied.

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The advice of Dr. J. R. Wait and Mr. P. A. Field in the preparation of the paper is acknowledged.

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ERRORS IN BRIDGE MEASUREMENTS

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SUMMARY.—Any linear network used as a bridge can be regarded as subject to three errors which are independent of the unknown and standard impedances (or admittances) being compared. These errors (which may be complex quantities having any argument) are: (a) a ratio error, (b) an error impedance in series with the standard, (c) an error admittance in parallel with the unknown. One of these errors can be reduced to zero by adjustment; the other two can be determined by measuring two known impedances (or admittances) of different orders of magnitude.

1. Introduction

At high frequencies, the presence of stray capacitances, etc., may cause the condition of balance to depart appreciably from the expected condition that the ratio of the unknown impedance to the standard impedance is a constant, usually real or purely imaginary, when the bridge is balanced. Here therefore we determine the conditions of balance ab initio in the general case, assuming only that various self and mutual impedances exist within the bridge network which are constant at any one frequency; we shall show that the bridge can be regarded as subject to three errors independent of each other and of the standard and unknown impedances being measured.

At first sight, it would appear that, in order to investigate bridge errors, we should make calculations based upon the known values of the elements composing the bridge. This procedure, however, involves two difficulties. At high frequencies there will be a number of elements such as stray capacitances and inductances which are not easily measured and, even if their values were completely known, the computation from such a complex arrangement of elements would be unnecessarily difficult. The second difficulty is that we wish to establish a general theorem, and to do so from particular circuits is difficult.

A new approach is required, analogous to 'ignoration of co-ordinates' in dynamics; we must refuse to find out anything we do not need to know. The relevant question is not "What causes this phenomenon?" but "What is its effect?". We should not ask "What are the electrical elements contained within this system?", but rather "What does this system do?".

In our particular problem, we do not need to know the elements forming the bridge network; we only need to know how the bridge behaves. We should think of the bridge network as enclosed within an unbreakable 'black box', accessible to the outside world only through four pairs of terminals; the only relevant data are those obtained from measurements at those terminals. The orderly tabulation of the results of such measurements leads us to the primary concept associated with this new 'black box' approach—the matrix.

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We first observe that the terminals have no significance unless current can flow between them. If we provide external connections between the terminals of a pair, these connections will determine the geometrical path of the resultant current and will, in fact, give arbitrary additional elements to the original network which is indeterminate. A possible way of counteracting this indeterminacy would be to short-circuit all terminal pairs; if this is done, the bridge network would be completely determined but also completely useless. Suppose now, however, that each short-circuiting link has a 'narrow' gap across which a voltage is injected from a zero-impedance source in series with a zero-impedance ammeter

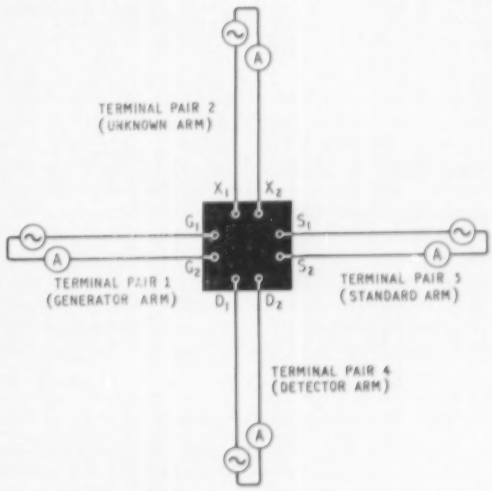

Fig. 1. *A bridge network with determinate current paths.*

as in Fig. 1. 'Narrow' means that the ratio of the gap width to the wavelength at the operating frequency is of the same order of magnitude as the ratio expressing the required precision. We are, under these circumstances, able to apply a voltage across any terminal pair and to measure the resulting current in the same or another

terminal pair without in any way altering the 'black box'. If now a voltage V_α is applied to the terminal pair T_α and the resulting current at the terminal pair T_j is $i_{\alpha j}$, we denote by $Y_{\alpha j}$ the quotient* $i_{\alpha j}/V_\alpha$, which is a constant of the network; $Y_{\alpha\alpha}$ is the current at terminal pair T_α per volt applied there, or the admittance at the terminal pair T_α, while $Y_{\alpha j}$ is similarly the mutual admittance between terminal pair T_α and terminal pair T_j, or the current measured at terminal pair T_j per volt applied at terminal pair T_α. By applying voltages in turn at each terminal pair and measuring the resultant $Y_{\alpha j}$, we determine the essential properties of the network, which may be conveniently tabulated in matrix form:

$$A = \begin{pmatrix} Y_{11} & Y_{12} & Y_{13} & Y_{14} \\ Y_{21} & Y_{22} & Y_{23} & Y_{24} \\ Y_{31} & Y_{32} & Y_{33} & Y_{34} \\ Y_{41} & Y_{42} & Y_{43} & Y_{44} \end{pmatrix} \quad .. \quad \text{(1a)}$$

If the principle of reciprocity applies (as it usually does) we shall have $Y_{jk} = Y_{kj}$, but this is not important from our present point of view. If the system is linear, the principle of superposition applies, so that if voltages V_1, V_2, V_3 and V_4 are applied simultaneously at the four terminal pairs

$$\left.\begin{aligned} i_1 &= Y_{11}V_1 + Y_{12}V_2 + Y_{13}V_3 + Y_{14}V_4 \\ i_2 &= Y_{21}V_1 + Y_{22}V_2 + Y_{23}V_3 + Y_{24}V_4 \\ i_3 &= Y_{31}V_1 + Y_{32}V_2 + Y_{33}V_3 + Y_{34}V_4 \\ i_4 &= Y_{41}V_1 + Y_{42}V_2 + Y_{43}V_3 + Y_{44}V_4 \end{aligned}\right\} \quad \text{(1b)}$$

At this stage we may, if we please, try to find some circuit which has the properties thus established and think in the familiar terms of the elements of that circuit. In fact, however, it is much easier to think entirely in terms of the Y_{ij} as the basic entities and resist the temptation to visualize them as circuit elements, since this visualization merely confuses the issue. It will be seen later that we ultimately need only three quantities to specify the behaviour of the bridge. These can be deduced from the Y_{ij} but, in practice, it is easier to measure them directly; the evaluation of the whole matrix (1a) is then unnecessary.

2. The Three Error Constants

The foregoing remarks apply generally to any network with four terminal pairs. If the network is used as a bridge, it has also the following special properties:

(i) We shall suppose that only terminal pair 1 is connected through a generator, so that V_1 and i_1 are arbitrary. The conditions for zero output from the detector terminals are independent of the magnitude of the input or the sensitivity of the detector. It follows that shunt or series elements associated with detector or source terminals have no effect on precision, but only on sensitivity. Further, the geometrical arrangement of the standards is determined a priori, so that the terminal pair where the unknown admittance is connected is the only one where the precise nature of the geometrical arrangement for the flow of current may be important.

(ii) If an unknown admittance Y_x is connected to terminal pair 2 and a standard admittance Y_s to terminal pair 3, we have, in addition to (1b)

$$i_2 = Y_xV_2; \quad i_3 = Y_sV_3 \quad .. \quad \text{(2)}$$

As we are only concerned with the conditions under which no current flows through the detector (assumed to have finite impedance Y_d), we have also

$$i_4 = Y_dV_4 = 0 \quad .. \quad .. \quad .. \quad \text{(3)}$$

The legitimacy of adding equations (2) and (3) to (1) is more fully discussed in the Appendix. Eliminating i_2 and i_3 by means of (2), we therefore have that, when no current flows through the detector,

$$\left.\begin{aligned} Y_{21}V_1 + (Y_{22} - Y_x)V_2 + Y_{23}V_3 &= 0 \\ Y_{31}V_1 + Y_{32}V_2 + (Y_{33} - Y_s)V_3 &= 0 \\ Y_{41}V_1 + Y_{42}V_2 + Y_{43}V_3 &= 0 \end{aligned}\right\} \quad .. \quad \text{(4)}$$

These equations are inconsistent unless

$$\begin{vmatrix} Y_{21} & Y_{22} - Y_x & Y_{23} \\ Y_{31} & Y_{32} & Y_{33} - Y_s \\ Y_{41} & Y_{42} & Y_{43} \end{vmatrix} = 0 \quad \text{(5)}$$

This determinant can be expanded in the form

$$Y_{41}\,Y_xY_s + \lambda\,Y_x + \mu\,Y_s + \nu = 0 \quad \text{(6)}$$

where

$$\lambda = Y_{31}Y_{43} - Y_{41}Y_{33}\,; \quad \mu = Y_{21}Y_{42} - Y_{41}Y_{22}$$

and

$$\nu = \begin{vmatrix} Y_{21} & Y_{22} & Y_{23} \\ Y_{31} & Y_{32} & Y_{33} \\ Y_{41} & Y_{42} & Y_{43} \end{vmatrix} \quad .. \quad .. \quad \text{(7)}$$

so that λ, μ, ν are constants of the network and independent of Y_x, Y_s. The importance of (6) is its form, bilinear in Y_x, Y_s.

Now consider the relation

$$Y_x + Y_e = \frac{\xi}{\dfrac{1}{Y_s} + Z_e} \quad .. \quad .. \quad \text{(8)}$$

It is the same as (6) (when cleared of fractions) provided we choose the, as yet unspecified, constants Y_e, Z_e and ξ so that

$$Y_{41}/Z_e = \lambda = \mu/(Z_eY_e - \xi) = \nu/Y_e \quad \text{(9)}$$

or

$$Z_e = Y_{41}/\lambda; \quad Y_e = \nu/\lambda; \quad \xi = (Y_{41}\nu - \mu\lambda)/\lambda^2 \quad \text{(10)}$$

The relations (6) and (8) specify that the network

* This assumes that all quantities involved are expressed in operational form.

behaves as if it were an ideal bridge with an error impedance Z_e in series with the standard and an error admittance Y_e in parallel with the unknown, the ratio of the bridge being ξ, where Z_e, Y_e and ξ are given by (10) and may be complex quantities having arguments of any value. It does not mean that the bridge network actually contains such elements, but that its errors can be corrected as if these elements were present.

If we have actually carried out the measurement of all the Y_{ij}, (10) specifies Z_e, Y_e and ξ completely. But, in order to determine the behaviour of the bridge, even measurement of the Y-matrix (1a) gives us more information than we need; it is sufficient to determine by measurement the quantities Y_e, Z_e and ξ on the assumption that a relation of the form (8) exists. For an admittance bridge, we first consider open-circuit conditions; that is to say, we have the standard side open-circuited ($Y_s = 0$) and provide an adjustable impedance in parallel with the unknown which is used to balance the bridge. When the bridge is balanced, the value of the adjustable admittance is $-Y_e$. We can, if we wish, note the value of Y_e thus obtained and then, by measurement on the bridge of two known impedances, obtain simultaneous equations for Z_e and ξ, but it is simpler to leave the adjustable impedance permanently at the value which secured balance on open circuit. By doing this we have, in effect, replaced the original bridge by a similar bridge free from the error impedance in parallel with the unknown. If the admittance $-Y_e$ has a negative real part, we shall find that we cannot balance the bridge on open circuit with our adjustable admittance unless a suitable fixed admittance is permanently incorporated in the bridge in parallel with the standard. This admittance does not need to be adjustable; its purpose is merely to make the effective value of $-Y_e$ have a positive real part. We assume that the adjustable admittance, and if necessary the fixed admittance, once determined, are added permanently to the structure of the bridge within the 'black box'; the relevant values of the elements of the matrix (1a) are henceforward those obtained after this change.

Having determined Y_e, or adjusted it to an effective value zero in the manner just described, we measure two known admittances, preferably of different orders of magnitude, with the bridge. (8) then gives us linear simultaneous equations for Z_e and ξ, and (8) can thereafter be used with these values of Z_e and ξ to determine any Y_x, given Y_s.

For an impedance bridge, the procedure is similar except that we first consider short-circuit conditions (Y_x infinite). An adjustable impedance is now required in series with the standard, and its value for balancing the bridge is $-Z_e$. A fixed impedance in series with the unknown may also be required if $-Z_e$ has a negative real part, so that the bridge cannot be balanced under short-circuit conditions otherwise. Having determined or removed Z_e, ξ and Y_e are determined by applying (8) to measurements of two known impedances of different orders of magnitude on the bridge; this gives linear simultaneous equations for ξ and Y_e.

We thus find that any admittance bridge can be regarded as subject to three errors which are independent of the standard and unknown impedances, namely

(a) an error impedance in series with the standard
(b) an error admittance in parallel with the unknown
(c) a ratio error.

The error (b) can be eliminated by adjustment, while (a) and (c) can be determined by measuring two known admittances, preferably of different orders of magnitude.

Similar remarks apply to impedance bridges, but now it is the error (a) which can be eliminated by adjustment.

Acknowledgment

The authors wish to thank the Chief Engineer of the British Broadcasting Corporation for permission to publish this paper.

APPENDIX

Assessing the Effect of Adding External Impedances Across the Terminals of a Network: Deduction of Thévenin's Theorem

Consider an arbitrary linear network A having terminal pairs T_j. Suppose a generator system G is connected to A so as to inject currents i_j into terminal pairs T_j. These currents are of the same frequency but supposed arbitrary in phase and amplitude; they are to be regarded as independent variables. In theory they should, therefore, be supplied from infinite-impedance sources. In practice, any generator may be considered as an infinite-impedance source shunted by a finite admittance, and these finite admittances Y_j shunted across the T_j would be such that the given network A could be regarded as equivalent to a network A' different from A but now connected to infinite-impedance sources. If we assume that the generators G are infinite-impedance sources, the currents i_j are to be regarded as injected into A without altering A or being altered by it.

If now the voltage across T_j is V_j then

$$V_j = \sum Z_{jr} i_r \qquad \text{(A1)}$$

The quantities Z_{jr} form the network impedance matrix and (A1) is in fact merely a statement that the system is linear.

We now connect across any one terminal pair T_x a new impedance Z_x.

We can, if we like, regard the network having this additional impedance as an entirely new network and obtain equations of the form (A1) ab initio with an

impedance matrix changed because of the addition of the new element Z_α.

Alternatively, the new voltages can be found by a method analogous to the principle of virtual work in dynamics. Suppose that the current in Z_α is i_α. Let the impedance Z_α be replaced by an additional infinite-impedance generator linked to the system G and delivering a current i_α into the terminals T_α.

We now have one more input current and the new voltages will be

$$V'_j = \sum Z_{jr} i_r + Z_{j\alpha} i_\alpha \quad .. \quad .. \quad .. \quad \text{(A2)}$$

for all terminal pairs including T_α.

Also $V'_\alpha = -Z_\alpha i_\alpha \quad .. \quad .. \quad .. \quad .. \quad \text{(A3)}$

The new current i_α does not differ in kind from any of the other i_j and, like them, it does not change A. These equations (A2) and (A3) determine all the V'_j. They are, in fact, a generalized form of the equations (1) and (2) used in the main text (with the parts played by current and voltage interchanged) and can clearly be generalized still further, for we could proceed as above if there had been any number m of external impedances Z_α instead of only one. If the network had n terminal pairs, we should obtain n linear equations like (A2) involving the currents i_r in the original network as well as the m currents like i_α in the external impedances, and we should also have m equations like (A3). The equations like (A3) could be used to replace each i_α by a multiple of V'_α, so that we should finally have n linear simultaneous equations for the V'_j in terms of the known currents i_r in the original network.

If in (A2) and (A3) we make Z_α tend to infinity, i_α will tend to zero and the voltage V'_α obtained across the terminals T_α is the open circuit voltage V_α so that

$$V_\alpha = \sum Z_{\alpha r} i_r \quad .. \quad .. \quad .. \quad .. \quad \text{(A4)}$$

Hence for terminal pair T_α, (A2) can be written

$$V'_\alpha = V_\alpha + Z_{\alpha\alpha} i_\alpha \quad .. \quad .. \quad .. \quad .. \quad \text{(A5)}$$

Combining (A3) and (A5)

$$(Z_\alpha + Z_{\alpha\alpha}) i_\alpha = -V_\alpha \quad .. \quad .. \quad .. \quad \text{(A6)}$$

If in (A6) we put $Z_\alpha = 0$, i_α becomes the short-circuit current $i_\alpha{}^*$ at T_α

and $Z_{\alpha\alpha} = -V_\alpha / i_\alpha{}^* \quad .. \quad .. \quad .. \quad .. \quad \text{(A7)}$

Also if all the i_j except i_α are reduced to zero

$$V''_\alpha = Z_{\alpha\alpha} i_\alpha$$

so that $Z_{\alpha\alpha}$ is the impedance of the network A from the terminals T_α.

Thus (A7) is in fact Thévenin's Theorem obtained by specialization of (A2) and (A3).

OPTIMUM *RC* FILTERS

For Separating Sinusoidal Signals from Noise

By J. W. R. Griffiths, B.Sc., A.M.I.E.E.

Introduction

IT is sometimes necessary in practice to separate a sinusoidal signal from unwanted fluctuations and, if very low frequencies are involved, normal band-pass filter techniques cannot be used. This problem arises in the detection of a noise source, such as a radio star, by interferometer techniques[1,2,3]. In brief, these consist of two receiving systems the outputs of which are either multiplied together or, in the older systems, added and rectified, the resultant in either case being passed through a low-pass filter. In the multiplier type, the d.c. output from the filter effectively measures the correlation between the outputs of the two receiving systems and, hence, the single point noise source whose signal at the two receiving systems is correlated, produces a d.c. output: the background and receiver noise, however, being uncorrelated in the two systems, produces a fluctuating output which can, to some extent, be reduced by lowering the cut-off frequency of the low-pass filter.

In the additive type a d.c. is present even when the signal is absent, and the presence of the signal is indicated by an increase in this d.c.; this means that in order to decide that a source has been detected, it is necessary to observe the background with and without the star present. This restriction does not apply to the multiplier type since d.c. is only present when a signal is being received.

If there is relative angular motion between the receiving system and the signal source, the signal output oscillates according to the interferometer pattern, so turning the d.c. output into a quasi-sinusoid and giving the signal a characteristic signature. This makes detection possible in the additive type, and considerably enhances detection in the multiplier type. The relative motion when searching for a radio star is conveniently achieved simply by the movement of the aerial system due to the earth's rotation. The frequency of the quasi-sinusoid sets a lower limit to the reduction of the background fluctuation by the low-pass filter, and it is with this point that this paper is mainly concerned.

In theoretical calculations of the minimum detectable input signal/noise ratio[2,3], it is usually assumed that the low-pass filter used is ideal; i.e., it is a low-pass filter which passes the wanted frequency without attenuation, but rejects all noise whose frequency components exceed this. However, since the frequency involved is extremely low—a typical figure is one cycle/hour—the realization of such a filter is impracticable, and normally an *RC* network of one or more sections is used. It is shown below that in such

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a case there is an optimum time constant for the *RC* network which is dependent on the number of sections but, even using this optimum, the signal/noise ratio is still inferior to that obtained with an ideal filter.

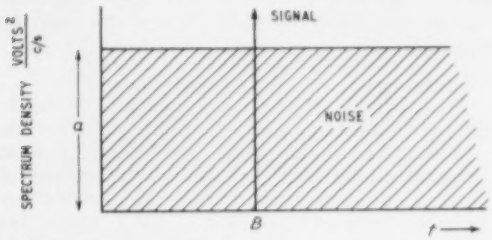

Fig. 1. *Spectrum assumed for noise and signal at the output of the system.*

Calculation of Optimum Conditions

Fig. 1 shows the spectrum assumed for the noise and signal at the output of the multiplier. The wanted frequency is assumed sinusoidal and the noise spectrum flat over the relevant frequency band. The noise spectrum is in fact triangular, but if the bandwidth of the receiving systems is much greater than B—as is usually the case—then the lower portion of the spectrum can be assumed uniform. The noise output from an ideal filter of cut-off frequency B is aB, a being the spectrum density of the noise in units of (voltage)^2 per c/s and, since the signal is unattenuated, the signal-to-noise power ratio

$$\frac{S}{N} = \frac{A^2}{aB} \quad A = \text{Signal amplitude (r.m.s.)}$$

For n identical *RC* circuits isolated from each other as in Fig. 2(a) the ratio of the output to the input voltage for any one frequency is given by

$$\frac{v_o}{v_i} = \frac{1}{(1 + jx)^n}$$

where $x = \dfrac{f}{f_o}$ and $f_o = \dfrac{1}{2\pi RC}$

$$\therefore \left|\frac{v_o}{v_i}\right|^2 = \frac{1}{(1 + x^2)^n}$$

Noise power in band $df = a.df$. [Unit resistance is assumed for convenience in working.]

Total noise power at output of filter

$$N = \int_0^\infty \frac{a.df}{(1 + x^2)^n}$$

Putting $x = \tan\theta$ we obtain

$$N = a\int_0^{\pi/2} (\cos\theta)^{2(n-1)}\, d\theta$$

The integrand can be expanded into a series containing one term which is a function of n only; the remaining terms contain a function of n and a cosine of an even multiple of θ and when integrated between 0 and $\pi/2$, contribute nothing to the final integral.

Hence we find the noise output

$$N = K.a.f_o$$

where $$K = \frac{\pi}{2} \cdot \frac{[2(n-1)]!}{2^{2(n-1)}[(n-1)!]^2}$$

$$\text{Signal output} = \frac{A}{\left\{1 + \left(\dfrac{B}{f_o}\right)^2\right\}^{n/2}}$$

$$\therefore \frac{S}{N} = \frac{A^2}{K.a.f_o.\left\{1 + \left(\dfrac{B}{f_o}\right)^2\right\}^n}$$

This can be shown to have a maximum when

$$\left(\frac{B}{f_o}\right)^2 = \frac{1}{2n-1}$$

and substituting we find

$$\left(\frac{S}{N}\right)_{opt} = K_o \frac{A^2}{a.B}$$

where $$K_o = \frac{\left(1 - \dfrac{1}{2n}\right)^n}{K}$$

Considering Fig. 2(b) in which the *RC* circuits are not isolated, we find for two sections,

$$\frac{v_o}{v_i} = \frac{1}{1 - x^2 + 3jx} \quad x \text{ being defined as before.}$$

$$\left|\frac{v_o}{v_i}\right|^2 = \frac{1}{1 + 7x^2 + x^4}$$

Total noise power $N = a.f_o \int_0^\infty \dfrac{dx}{1 + 7x^2 + x^4}$

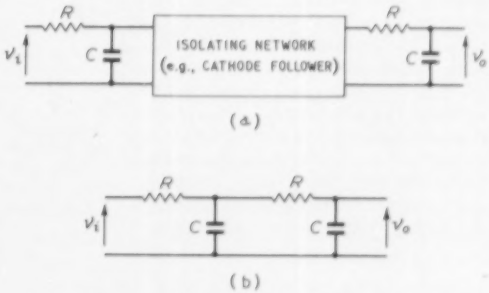

Fig. 2. *RC filter networks.*

By the substitution $y = \dfrac{1}{x}$ it can be shown that

$$N = a.f_o \int_0^\infty \frac{dx}{\dfrac{1}{x^2} + 7 + x^2}$$

and adding gives

$$2N = a.f_o \int_0^\infty \frac{\left(1 + \frac{1}{x^2}\right) dx}{\frac{1}{x^2} + 7 + x^2}$$

Putting $u = x - \frac{1}{x}$ we obtain

$$N = a.f_0 \int_0^\infty \frac{du}{u^2 + 9}$$

$$= \frac{a.f_o.\pi}{6}$$

$$\therefore \frac{S}{N} = \frac{6A^2}{\pi.a.f_o\left\{1 + 7\left(\frac{B}{f_o}\right)^2 + \left(\frac{B}{f_o}\right)^4\right\}}$$

which is a maximum when

$$\frac{B}{f_o} \simeq \frac{1}{\sqrt{7{\cdot}4}}$$

$$\therefore \left(\frac{S}{N}\right)_{opt} = 0{\cdot}36\frac{A^2}{aB}$$

Discussion

Table 1 summarizes the results. Column 4 assumes that there is a square-law relationship between the input and output S/N ratios, an assumption valid for all input S/N ratios in the case of the multiplier types, but only valid for S/N ratios much less than unity in the additive types.

In making a decision as to the best RC filter for any particular system an important point for consideration is the question of the total RC product of the network. The resistance R is limited by leakage currents, and therefore the value of the product RC determines the required value of the capacitance C, and so the size and weight of the filter. As very long time constants are required, this becomes a limiting factor to the design of the filter.

When two isolated sections are used under optimum conditions, each section has an RC product only $1/\sqrt{3}$ of the single section, so that the total is increased to $2/\sqrt{3}$. In the case of cascaded sections each RC product is reduced to $1/\sqrt{7{\cdot}4}$ so that the total RC product $2/\sqrt{7{\cdot}4}$ is less than the original.

Conclusions

When RC circuits are used to reduce the background fluctuations and so facilitate the detection of a weak sinusoidal signal, there is an optimum time constant which depends on the number of sections, and whether or not the sections are isolated from one another. Even using the optimum values, an RC filter of any number of sections still falls short of the performance of the ideal filter.

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TABLE 1

1 Filter	2 $(S/N)_{opt}$ (power ratio)	3 Loss compared with ideal filter (decibels)	4 Increase of input S/N to give same output S/N as ideal filter (decibels)	5 Optimum conditions
Ideal	$1{\cdot}00\frac{A^2}{a.B}$	0	0	Cut-off of filter $= B$
One RC section	$0{\cdot}32\frac{A^2}{a.B}$	5	2·5	$f_o = B$
Two isolated RC sections	$0{\cdot}41\frac{A^2}{a.B}$	4	2	$f_o = \sqrt{3}\,B$
A large number of isolated RC sections	$0{\cdot}51\frac{A^2}{a.B}$	3	1·5	$f_o = \sqrt{2n}\,B$
Two cascaded RC sections	$0{\cdot}36\frac{A^2}{a.B}$	4·4	2·2	$f_o = \sqrt{7{\cdot}4}\,B$

FLUCTUATION NOISE

Direct Measurement at Very Low Frequencies

By F. J. Hyde, M.Sc., A.M.I.E.E.

(Official communication from D.S.I.R. Radio Research Station, Slough)

SUMMARY.—A method of measuring the current noise generated in temperature-sensitive solid-state devices, in narrow bandwidths about frequencies chosen between 5×10^{-3} and 10^{-1} c/s is described. The central feature of the equipment is a resistance-capacitance coupled amplifier of the type originally described by Schneider (1945). Spurious electrical fluctuations arising from the circuitry and thermal fluctuations of the ambient oil surrounding the specimen were shown not to cause any error. Specification of the noise necessitates determination of the impedance of the specimen at the frequencies in question: this may be derived from measurements of the limiting small-signal resistances of the specimen at very low and very high frequencies and its thermal time-constant.

1. Introduction

IT has been of interest to measure the spectral density of fluctuation noise, at very low frequencies, associated with current-carrying germanium devices, as a means of investigating the fundamental noise-generation process. One method used elsewhere (Rollin and Templeton 1953, 1954) was to record the noise in a broad bandwidth on a magnetic tape running at slow speed. This was then played back at normal speed and selective amplification of the recorded noise effected in the audio-frequency range. From the ratio of recording to play-back speed the noise density was referred back to the appropriate low frequencies at which it was generated. By this means it proved possible to measure noise down to frequencies as low as $2{\cdot}5 \times 10^{-4}$ c/s.

With the equipment to be described here, noise has been measured *directly* in a narrow bandwidth at frequencies down to 5×10^{-3} c/s, being displayed as a pen recording on a chart. Measurements at lower frequencies could be made if sufficient recording time were allowed. The essential feature of the equipment is an amplifier, resonant at the frequencies at which recordings are to be made, with a Q-factor of about eight for each resonance. The type of resonant system used was originally described by Schneider (1945), although it is believed that the present is the lowest frequency range of application of the system so far reported.

2. Resonant Amplifier for Very Low Frequencies

A schematic diagram of the selective amplifier is shown in Fig. 1. The transmission is given by

$$\frac{e_o}{e_i} \propto \frac{1}{1 + jQ\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)} \quad .. \quad .. \quad (1)$$

as for an ideal tuned circuit. The angular resonant frequency is given by

$$\omega_0 = 1/(mR_1C_1R_2C_2)^{\frac{1}{2}} \quad .. \quad .. \quad (2)$$

and the Q-factor by

$$Q = \frac{m_1^{\frac{1}{2}}}{g}\left(\frac{R_2C_2}{R_1C_1}\right)^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad (3)$$

if the leakage resistances r_1 and r_2 of C_1 and C_2 can be ignored; i.e., if $\omega_0 C_1 r_1$, $\omega_0 C_2 r_2 > 10$ say. Here m is the aperiodic gain of the amplifier with the feedback capacitor C_2 removed,

$$m_1 = \left(m + 1 + \frac{R_3}{R_2} + \frac{R_3}{R_1}\right),$$

$$g = \left(1 + \frac{R_2}{R_1} + \frac{R_2 + R_3}{R_1}\cdot\frac{C_2}{C_1}\right)$$

and R_3 is the output resistance of the amplifier and is equal to the parallel combination of the anode load of V_3 (see Fig. 2) and the slope resistance of V_3.

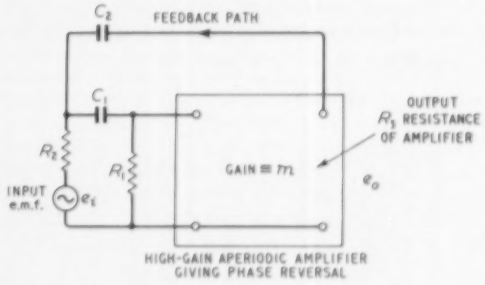

Fig. 1. *Schematic circuit of selective amplifier.*

The Q-factor may be maximized for a given gain m by a suitable choice of ratios C_2/C_1 and R_2/R_1 with respect to R_3/R_1. These are $C_2/C_1 = (R_1/R_3)^{\frac{1}{2}} = R_1/R_2$. It will be noted that the resonant frequency is inversely proportional to the square root of the aperiodic gain m of the amplifier, so that very low resonant frequencies can be obtained with quite ordinary values of R_1, R_2, C_1, C_2. The optimum value of Q is

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$$Q_{opt} = \frac{m_1^{\frac{1}{2}}}{2\left[1 + \left(\frac{R_3}{R_1}\right)^{\frac{1}{2}}\right]} \quad .. \quad .. \quad (4)$$

A detailed circuit diagram of the amplifier is given in Fig. 2. Valves 1–3 form a direct-coupled amplifier giving the required phase reversal with a voltage gain m of about 1000. Bias voltages are supplied mainly by potentiometers between h.t. and earth. By ganged switching of C_2 and C_1, maintaining their ratio approximately constant to preserve optimum Q, five resonant frequencies ($f_0 = \omega_0/2\pi$) were obtained as shown in Table 1.

TABLE 1

Switch Position	1	2	3	4	5	6
C_2 μF	2	1	0·5	0·25	0·1	–
C_1 μF	1	0·5	0·25	0·1	0·05	1
f_0 c/s (approx.)	0·005	0·01	0·02	0·04	0·10	Broad Band

The 0·001-μF capacitor across the final anode load suppresses oscillation at high frequencies. It was found necessary to use a separate stabilized h.t. supply for the cathode follower V_4, to avoid instability at low frequencies. Decoupling is not a practicable proposition at the low frequencies we are concerned with here. The heaters were d.c. supplied at 6·3 volts from two 6-volt accumulators with barretter stabilization. When the unit was preceded by a commercial d.c. amplifier, flicker noise was not troublesome at full gain, and drift of the output current through the pen recorder was not more than 0·1 mA in eight hours, which was the longest continuous recording period used. The resonant frequency may be varied slightly by changing the heater voltage of the valves, thereby altering the gain m.

3. Parameters of the Amplifier

In order that a measured noise current output may be referred to a noise voltage at the input of the amplifier, it is necessary to know the transfer conductance K of the unit at the resonant frequency f_0, and also the noise bandwidth $\frac{\pi}{2} \cdot \frac{f_0}{Q}$. A convenient way to determine these parameters is to record the transient response of the system to a unit step of voltage applied at the input terminals. The response of the amplifying section can be shown, with negligible approximation if $Q > 5$ say, to be

$$V(t) = \frac{G_0}{Q} \exp.\left(-\frac{\omega_0}{2Q} t\right) \sin \omega_0 t \quad .. \quad (5)$$

where $V(t)$ is the voltage developed at the anode of V_3, and $G_0 = m/g$ is the gain of the amplifier at resonance. If it is assumed that the pen-recorder response may be specified by a single time-constant $\tau_R < 2Q/\omega_0$ then it follows that the overall transient response is

$$I(t) = \frac{K}{Q} \exp.\left(-\frac{\omega_0}{2Q} t\right) \sin \omega_0 t \quad .. \quad (6)$$

where $K = G_0 K_c$, K_c being the transfer conductance of the output cathode-follower. The attribution of a single time-constant to the pen-recorder is a reasonable approximation if about 6 volts r.m.s. of 50-c/s excitation from a high-impedance source is superimposed on the

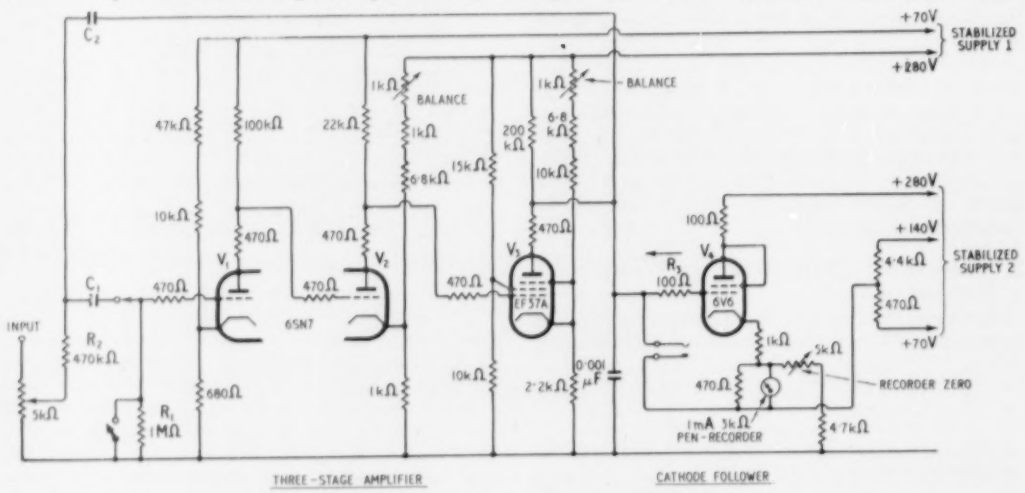
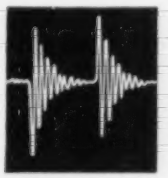


Fig. 2. *Circuit of very-low-frequency resonant amplifier.*

low-frequency noise applied to the pen-recorder movement, so that pen–paper friction is minimized: τ_R was about 0·1 sec for the recorder used here. A typical transient record resulting from a step of 0·279 volt applied at $t = 0$, and a plot of the dependence of the amplitude of the peaks of the transient waveform on the number of periods is shown in Fig. 3. The resonant frequency is determined by timing periods with a stop-watch.



TRANSIENT RESPONSE TO UNIT STEP VOLTAGE

AT INPUT TERMINALS $I(t) = \frac{K}{Q}$ EXP. $\left(-\frac{\omega_0}{2Q}t\right)$ SIN $\omega_0 t$

ω_0 = RESONANT FREQUENCY $(= 2\pi f_0)$
K = TRANSFER CONDUCTANCE

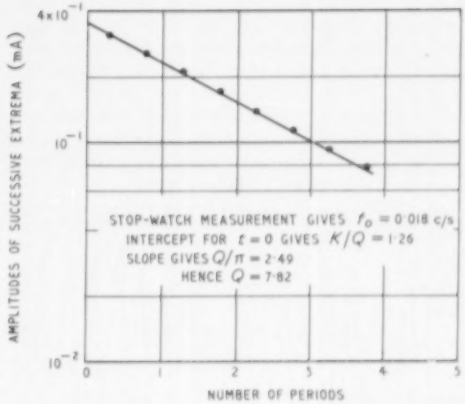
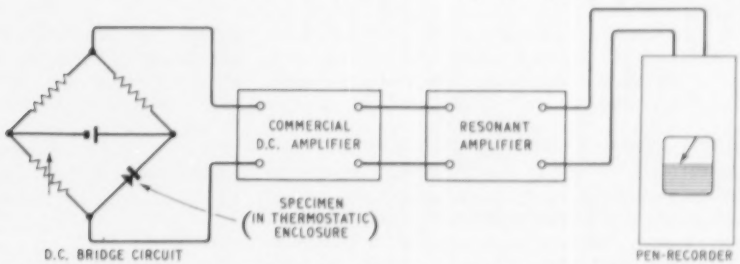


Fig. 3 (above). *Typical transient record and analysis.*

Fig. 4 (right). *Schematic circuit for measurement of noise.*



4. Complete Measuring Circuit for Noise

This is shown schematically in Fig. 4. The specimen under examination, shown as a rectifier biased in the reverse direction, is immersed in an oil-bath, the temperature of which is controlled to an accuracy of about $\pm$ 0·001 °C. It comprises one arm of a d.c. bridge circuit, the other arms of which are wire-wound resistances generating only thermal noise, even when carrying current. The bridge is adjusted so that the required direct current flows through the specimen and there is no direct voltage output at the bridge terminals.

The excess noise generated in the specimen, when current is passed through it, does, however, produce a *noise* voltage at the bridge output terminals (the total thermal noise at the bridge output is always very much smaller than the excess noise generated in the specimen at the low frequencies we are concerned with here). This is amplified by a commercial d.c. amplifier having a cut-off frequency of 10 c/s and a maximum sensitivity of 25 volts output for 1 mV input across 1000 ohms. The amplified broad-band noise is then further amplified selectively by the resonant amplifier, and the noise output of this is applied to the pen recorder. A section of a typical noise record for $f_0 = 0·01$ c/s, noise bandwidth 0·002 c/s and paper speed of twelve inches per hour is shown in Fig. 5.

5. Analysis of the Noise Records

It was shown by Rice (1945) that the envelope curve $A(t)$ of a noise current contained in a narrow band of frequencies Δf centred on f_0 has a Rayleigh distribution of amplitude: that is, the probability of the instantaneous amplitude of the envelope lying between A and $A + dA$ is

$$p(A)dA = \frac{A}{< \Delta I^2 >} \exp.\left(\frac{-A^2}{2 < \Delta I^2 >}\right) dA \qquad (7)$$

where $< \Delta I^2 >$ is the mean-square value of the noise current in the bandwidth Δf and is defined by

$$< \Delta I^2 > = \lim_{T \to \infty} \frac{1}{T} \int_0^T i^2(t)dt$$

$i(t)$ is the instantaneous value of the noise current, also in the bandwidth Δf. The present noise records have been analysed on the basis of this distribution.

In practice, the noise was recorded for sufficient time to enable about 250 complete waves to be obtained. For our conditions this meant that the probable error of determination of the true r.m.s. value of the noise current was not greater than 10%. This figure is derived as follows. If the 250 waves were all independent then there would be this number of data from which to determine the r.m.s. value of the wave. In fact there is correlation between adjacent waves arising from the resonant property of the selective circuit. For Qs of about eight the number of independent data is reduced to about 100. Using the statistical result that the standard deviation of the error of determination of the standard deviation of a distribution is approximately equal to the reciprocal of the square root of the number of data, we find this to be 1/10th or 10%.

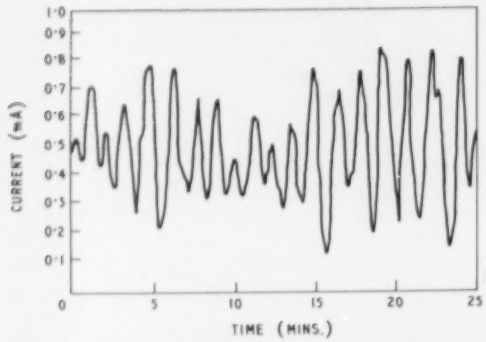


Fig. 5. *Section of a typical noise record at* 0·01 *c/s; noise bandwidth* 0·002 *c/s.*

A convenient plot, which was used to determine the r.m.s. value from the data, was the logarithm of the ratio of the number of readings N for which $(2A)^2$ exceeded the abscissa $\{(2A)^2\}$ to the total number of readings N_0. This leads to a straight line of slope $-1/8 < \Delta I^2 >$ as may be seen from equation (8).

$$N = N_0 \int_A^\infty P(A)dA$$

$$= N_0 \exp. - \{(2A)^2/8 < \Delta I^2 >\} \ldots \quad \ldots \quad (8)$$

A plot of this type is shown in Fig. 6.

When the mean square value $< \Delta I^2 >$ of the output noise-current wave had been obtained, it remained to determine from this the equivalent noise-current generator in shunt with the terminals of the specimen, by dividing by the overall current gain, which included the attenuation between the terminals of the specimen and the terminals of the d.c. amplifier (see next section). From a knowledge of the noise bandwidth of the system, $\Delta f = \pi f_0/2Q$, one could convert $< \Delta I^2 >$ to the spectral density of mean-square current, $< \Delta I^2 > /\Delta f$ for plotting purposes.

6. Impedance of the Specimen

To calculate the attenuation constant for the noise current between the terminals of the specimen and those of the d.c. amplifier, it is necessary to know the small-signal impedance of the specimen at its working point. Direct measurement of this, using conventional a.c. techniques, would have been extremely tedious even if a sine-wave generator for frequencies below 0·1 c/s, had been available.

The impedance at these frequencies may be inferred, however, from a knowledge of the thermal time constant τ of the noise-generating region of the device, the small signal a.c. resistance R_∞ measured at frequencies $\gg 1/\tau$, and R_0 the slope resistance of the static direct-current/voltage characteristic. It is implicit in this statement that any dispersive effects in the impedance in the frequency range considered here are to be attributed to direct thermal effects. Although long relaxation times, such that their reciprocals lie within this frequency range, are postulated to explain the spectral distribution of the noise intensity at low frequencies, it is not considered that the physical processes with which they are associated have a macroscopic effect on the impedance of the specimen.

Depending on whether the thermal time constant is greater or less than the reciprocal of the measuring frequency, either the small-signal slope resistance of the isothermal characteristic or of the static characteristic respectively is required. If the reciprocal of the thermal time constant lies in our frequency range then the locus of the small-signal current-voltage relationship about the working point becomes elliptical.

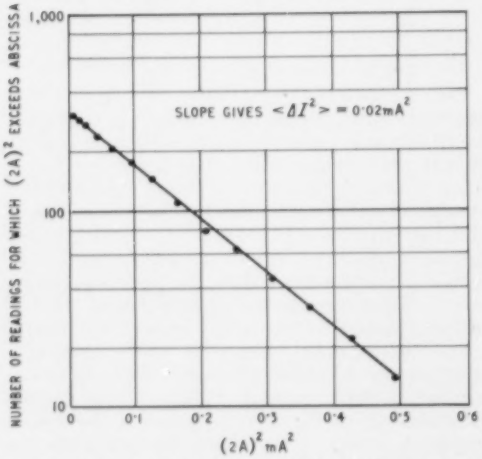


Fig. 6. *Plot of distribution of peak-to-peak amplitudes* $(2A)$ *of waves from which the mean square value* $< \Delta I^2 >$ *of the wave-train is deduced.*

The thermal time constant referred to is that of the noise-generating region. In germanium devices, in which the resistance and noise arise mainly from potential barriers so that power dissipation is extremely localized, the effective thermal time constant is very short, $< 10^{-6}$ sec (say). We are then concerned with the slope resistance of the static characteristic at our very low frequencies and this is taken to be the same as the a.c. resistance measured at a few cycles per second. For small power loadings, so that self-heating is negligible, the isothermal and static characteristics are coincident.

On the other hand, when the specimen under test is a filament, so that its resistance originates in the bulk, as will be the case if the end-connections are made substantially non-rectifying, then the time constant is much longer. The isothermal and static characteristics of a filament are not coincident for the relatively large power dissipations which sometimes ensue when large direct current is passed.

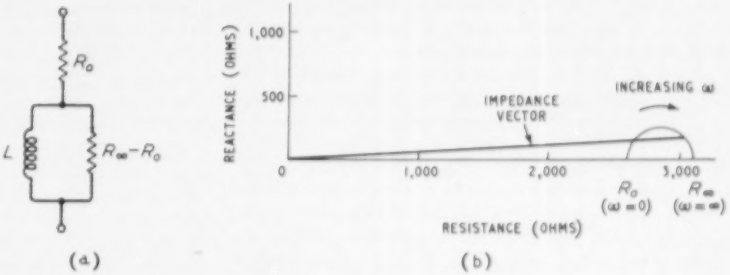


Fig. 7. Equivalent circuit (a) and locus of impedance vector (b) for a filament for which $R_\infty > |R_0|$: $L = 1040$ H.

The general theory of the impedance of temperature-sensitive circuit elements (Burgess 1955) may be adapted to show that if the quantities R_∞, R_0 and τ of a specimen are known, then the impedance at any frequency is given by

$$Z(\omega) = R_\infty - (R_\infty - R_0)/\{1 + j\omega\tau(R_\infty + R_0)/2R_\infty\} \quad .. \quad .. \quad (9)$$

for the special case in which the isothermal current-voltage characteristics are linear. R_∞ and R_0 are readily measured. The following technique has been used to determine τ. The filament forms one arm of a d.c. bridge circuit, the other arms of which are wire-wound resistances having small temperature coefficients of resistance. The bridge is balanced so that the current flowing through the filament is not large enough to cause measurable self-heating. By increasing the voltage applied to the bridge for a short time, the temperature of the filament may be raised by about 1°C (say) while the temperature of the ambient, which is usually oil, is substantially unaffected. If the bridge voltage is then reduced to its original value, the resistance of the filament will change at a rate determined solely by its thermal time constant which may be deduced from the rate of change of the out-of-balance waveform of the bridge, which is conveniently recorded photographically.

The actual mechanism of heat loss is a mixture of conduction through the leads and conduction and convection through the surrounding oil, so that the process cannot exactly be described by a single time constant. A good fit in terms of a single time constant is nevertheless obtained in practice.

In Fig. 7 are shown the equivalent circuit which may be derived from equation (9) and the locus of the impedance vector for a filament in which sufficient power was dissipated to raise its temperature, and for which $R_\infty = 3100\ \Omega$, $R_0 = 2600\ \Omega$, $\tau = 2{\cdot}3$ sec, and $L = \tau\,(R_\infty^2 - R_0^2)/2R_\infty = 1040$ H.

7. Sources of Error

It was necessary to ensure that the noise measured was directly attributable to the fundamental noise generation process in the specimen. Other causes of fluctuating output from the system were therefore carefully investigated. Undesirable fluctuations which may have been sufficiently great in magnitude to cause error could have originated in (a) the dry batteries supplying current to the bridge circuit of which the specimen was a member, (b) the amplifiers, (c) the thermostatic enclosure, as temperature variations.

Noise generated in the batteries would have produced a noise output at the bridge terminals if the a.c. and d.c. impedances of the specimen were different (as was usually the case.) The battery used to feed the bridge was loaded so that currents flowed equal to those drawn from it during normal operation: a similar loaded battery was connected in series with it, in opposition, and the combined output voltage adjusted to zero. The batteries were separated from each other physically, and electrically and thermally screened. When connected to the noise-measuring equipment operating at full gain no increase in output was detected. This test also showed that noise of the noise-measuring equipment was also negligible. At full gain, circuit noise was just detectable as a fluctuating output current.

Temperature fluctuations in the oil-bath were minimized by a suitable design. A quart thermos flask containing transformer oil was surrounded

by a metal cylinder, with the intervening space of about $\frac{1}{2}$ inch all round filled with glass wool. This cylinder acted as an electrostatic screen and also as a uniform temperature surface. An enamelled resistance wire was wound on the cylinder: power dissipated in this was the sole source of heat. The cylinder was itself enclosed in a wooden box, the intervening space again being filled with glass wool. A control thermistor sampled the temperature as close to the surface of the metal cylinder as its envelope would allow. It formed one arm of a resistance bridge fed with 50-c/s a.c. The out-of-balance voltage from the bridge, after amplification, was used to control the striking of a thyratron valve which in turn controlled the supply of electrical power to the heater winding. The parameters were such that at the location of the thermistor the temperature varied by $\pm 0{\cdot}02$°C about that value for which the control thermistor bridge was balanced.

The waves of heat passing through the cylinder into the oil were considerably attenuated and, in consequence, the internal temperature of the oil-bath fluctuated less than that of the control thermistor. An assessment of the internal fluctuations was made by replacing the specimen in the enclosure by a thermistor and attempting to measure the noise output arising from temperature-dependent resistance variation. None was detected, showing that temperature fluctuations were unimportant. An upper limit for the temperature fluctuations can be obtained as follows.

If α is the temperature coefficient of the resistance R of this thermistor when fed under practically constant voltage conditions, then a change of temperature ΔT will give rise to a change in current through the thermistor $\Delta I = I\alpha\Delta T$ and $< \Delta I^2 > = I^2\alpha^2 < \Delta T >^2 = \int_0^\infty S(f)_{i,T}$. $S(f)_{i,T}$, the spectral density of the mean square current fluctuation, is then given by

$$S(f)_{i,T} = 4I^2\alpha^2\{\tau/(1 + \omega^2\tau^2)\} < \Delta T^2 >$$

where τ is the thermal time constant of the thermistor (c.f. Schottky 1926). The minimum detectable spectral density of mean-square noise current at $f_0 = 0{\cdot}01$ c/s was 10^{-17} amp^2 sec; for the thermistor used, $I = 10^{-3}$ amp, $\tau = 1{\cdot}4$ sec, $\alpha \approx 0{\cdot}04$ deg^{-1}. Hence

$$< \Delta T^2 > \ll 2 \times 10^{-5} \text{ deg}^2$$

8. Conclusions

The paper has demonstrated that direct measurement of current noise in semiconductors in the frequency range 5×10^{-3} to 10^{-1} c/s is possible. The major source of inaccuracy at the lowest frequencies has been shown to arise from the practical consideration of length of recording time. The Q-factor of about eight chosen is a reasonable compromise at these frequencies: if a larger value is chosen, a correspondingly longer recording time must be allowed to achieve the same accuracy of measurement because of the longer persistence of strong correlation between adjacent waves of the noise record: if a smaller value is chosen then, with a frequency spectrum varying inversely as the frequency, the actual location of the noise spectral density is uncertain.

The alternative method of recording broad-band low-frequency noise on a magnetic tape has the attraction that one actual noise record contains data on all frequencies in the pass-band of the recording equipment. Sensitivity may be reduced by the noisiness of the tape.

The actual noise measurements made with the equipment described here are published elsewhere, since they are best regarded in their context; that is, as details of the more complete noise spectra extending from 5×10^{-3} c/s to several Mc/s.

Acknowledgments

The author wishes to acknowledge the assistance of Messrs. H. C. Bevan and R. E. Taylor with the experimental work.

The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of Radio Research of the Department of Scientific and Industrial Research.

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PROFESSOR G. W. O. HOWE

The Council of the British Institution of Radio Engineers has elected Professor G. W. O. Howe an Honorary Member of the Institution "in recognition of his contributions to the advancement of radio science, and in particular for his long series of notable articles in *Wireless Engineer*".

ELECTRONIC INSTRUMENTS EXHIBITION

An exhibition of electronic instruments will be staged by E.M.I. Electronics Ltd., at the Royal Hotel, Woburn Place, London, W.C.1, on November 28th–30th, 10 a.m. to 6 p.m. Tickets may be obtained free of charge from E.M.I. Electronics Ltd., Hayes, Middlesex, on written application.

NEW BOOKS

Analysis of Electric Circuits

By William H. Middendorf. Pp. 306 + xiv. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 48s.

This American book starts with the basic laws and definitions of electricity and then proceeds rapidly to alternating current and voltage, notation, and the algebra of complex numbers. It treats only alternating current and voltage in the steady-state. Seven chapters cover impedance, circuit parameters, calculation of power and series and parallel circuits.

A chapter on locus diagrams is then included and followed by a treatment of resonance. Circuits having more than one source of current or voltage are discussed and Part 1 ends with Chapter 16 on magnetic-field coupling in which transformers are treated.

Part 2 has only two chapters, on non-sinusoidal waves and network simplifications and theorems. Part 3 again has two chapters, in which polyphase circuits and communication circuits are discussed.

The author sticks strictly to the 'electric' of his title and makes no mention of valve circuits or of any electronic matters. It is a good point that he defines very clearly early on in the book the conventions that he adopts for the analyses and for the nomenclature. Not everyone will like his choice, but at least one always knows what he means.

The form of treatment is conventional and is carried out to a great extent by complex algebra. The polar form of representation is, however, used rather more than usual. W.T.C.

Frequency Modulation Engineering (Second Edition)

By Christopher E. Tibbs, M.I.E.E., M.Brit.I.R.E. and G. G. Johnstone, B.Sc. Pp. 435 + xii. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 45s.

The first edition of this book appeared in 1947 and was due to the first author alone. The revision which has resulted in this second edition has been carried out mainly by the second author.

There are eleven chapters, of which the first four introduce the subject and deal with the theoretical aspects, including interference and its suppression and noise structure. Propagation, which means v.h.f. propagation, is then discussed.

Chapter 6 covers aerials and feeders, the emphasis being placed on transmitting rather than receiving types. Transmitters for f.m. are treated in Chapter 7. Limiters, discriminators and f.m. receivers are dealt with in the next two chapters, and the book concludes with chapters on measurements and the practical uses of f.m.

Each chapter includes a list of selected references. The treatment is, to a large extent, explanatory and the mathematics, which is not extensive, is of a fairly simple nature. W.T.C.

Abacs or Nomograms

By A. Giet. Translated and revised by J. W. Head, M.A. (Cantab.) and H. D. Phippen, M.A. (Edin.), B.Sc. (Lond.). Pp. 225 + ix. Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 35s.

This book explains how to construct nomograms, starting with simple graphs and going on by easy stages to the multiple parallel-index-line alignment chart. Many examples of different kinds of chart are included. The book has been translated from the original French.

Hi-Fi from Microphone to Ear

By G. Slot. Pp. 180. Philips Technical Library. Distributed in the U.K. by Cleaver-Hume Press Ltd., 31 Wright's Lane, Kensington, London, W.8. Price 17s. 6d.

Introduction to Distributed Amplification

By Harry Stockman, S.D. Pp. 240. S.E.R. Co., 543 Lexington Street, Waltham, Mass., U.S.A. Price $2.90.

Radio Research 1955

Pp. 56. Reports of the Radio Research Board and the Director of Radio Research. Published for Department of Scientific and Industrial Research by H.M.S.O., York House, Kingsway, London, W.C.2. Price 3s. 6d.

ASLIB Year Book 1956–1957

Pp. 172 + xv. ASLIB, 4 Palace Gate, London, W.8. Price 12s. 6d.

TV Repair Questions and Answers: Vol. 5. Sound and Low Voltage Circuits

By Sidney Platt. Pp. 120. John F. Rider Publisher Inc., 480 Canal Street, New York 13, N.Y., U.S.A. Price $2.10.

Co-Operative Electrical Research

Journal of the Electrical Research Association, Vol. 1, No. 1, July 1956. Pp. 40. The Electrical Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 2s. 6d.

Les Antennes

By L. Thourel. Pp. 440. Dunod, 92 Rue Bonaparte, Paris 6e. Price 4,800 F.

Intended to supplement theoretical works in aerials, this book uses the minimum amount of mathematics consistent with a proper understanding of the ideas put forward. Calculations are backed up with experimental results, and the degree of practical validity of formulae indicated. Contents include chapters on medium-wave transmission; half-wave aerials for metric and decimetric waves; short-wave transmission; end-fire aerials; wideband omnidirectional v.h.f. and u.h.f. aerials; radiation from openings; horns; paraboloids; aerials with wide beam-widths in one plane and narrow beam-widths in another; slot aerials; lenses and frame aerials. An appendix deals with the Smith Chart.

The Distribution of Radio Brightness on the Solar Disk: Interstellar Hydrogen

International Scientific Radio Union (U.R.S.I.) Special Reports Nos. 4 and 5, in one cover. Pp. 72. General Secretariat of U.R.S.I., 42 rue des Minimes, Brussels, Belgium. Price 14s. 6d. (including postage).

Radio and Electronic Components Vol. 2—Variable Resistors and Potentiometers

By G. W. A. Dummer, M.B.E., M.I.E.E. Pp. 176 + xiii. Sir Isaac Pitman & Sons Ltd., Parker Street, Kingsway, London, W.C.2. Price 30s.

The Robertson Guide to Screw Thread Forms

Leaflet from W. H. A. Robertson & Co. Ltd., Lynton Works, Bedford.

BRITISH STANDARDS

Methods of Testing Vulcanized Rubber

B.S. 903. Part C3: 1956. Price 2s. 6d.

Memorandum on the Design of Electrical Apparatus having Double Insulation

B.S. 2754: 1956. Price 4s.

Varnished Cotton Cloth Sheet and Tape for Electrical Purposes
B.S. 419: 1956. Price 5s.

Terminology of Internal Defects in Castings as Revealed by Radiography
B.S. 2737: 1956. Price 12s. 6d.

Basic Characteristics of Radio Service Selection and Intercommunication Systems for Civil Aircraft
B.S. R.2: 1956. Price 3s.

Thin Vulcanized Fibre Sheets (Including Leatheroid) for Electrical Purposes
B.S. 2768: 1956. Price 4s.

These publications can be obtained from the British Standards Institution, 2 Park Street, London, W.1.

NATIONAL BUREAU OF STANDARDS

Protection against Betatron-Synchrotron Radiation up to 100 Million Electron Volts
National Bureau of Standards Handbook 55. Pp. 52. Price 25 cents.

Units and Systems of Weights and Measures
By Lewis V. Judson. Pp. 29. Circular No. 570. Price 25 cents.

Electron Physics Tables
By L. Marton, C. Marton and W. G. Hall. Pp. 83. Circular No. 571. Price 50 cents.

Amplitude and Phase Curves for Ground-Wave Propagation in the Band 200 c/s to 500 kc/s
By James R. Wait and H. Herbert Howe. Pp. 17. Circular No. 574. Price 20 cents.

These publications can be obtained from National Bureau of Standards, U.S. Department of Commerce, Washington 25, D.C., U.S.A.

MEETINGS

I.E.E.

14th November. "Frequency Diversity in the Reception of Selectively Fading Binary Frequency-Modulated Signals with special reference to Long-Distance Radiotelegraphy", by J. W. Allnatt, B.Sc.(Eng.), E. D. J. Jones and H. B. Law, B.Sc.Tech. "An Investigation of the Spectra of Binary Frequency-Modulated Signals with Various Build-up Waveforms", by J. W. Allnatt, B.Sc.(Eng.) and E. D. J. Jones. "An Improved Fading Machine", by H. B. Law, B.Sc.Tech., F. J. Lee, F. A. W. Levett and R. C. Looser, B.Sc.(Eng.). "The Detectability of Fading Radiotelegraph Signals in Noise", by H. B. Law, B.Sc.Tech. "The Signal/Noise Performance Rating of Receivers for Long-Distance Synchronous Radiotelegraph Systems using Frequency Modulation", by H. B. Law, B.Sc.Tech.

22nd November. "The Presentation and Demonstration of the Theory of Semi-Conductors to Students of Electrical Engineering." Discussion to be opened by P. Godfrey, B.Sc.(Eng.) at 6 o'clock.

29th November. "Differences of Opinion about Dimensions", Professor R. O. Kapp, B.Sc.(Eng.).

3rd December. "Electronics and Automation—The Use of Nucleonic Devices", by Denis Taylor, M.Sc., Ph.D.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, at 5.30 p.m., except where otherwise stated.

Brit.I.R.E.

28th November. "Colour Television", by G. N. Patchett, Ph.D., to be held at 6.30 at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.

The Television Society

22nd November. "Alternatives to the N.T.S.C. Colour System", by Dr. E. L. C. White, M.A., to be held at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2, at 7 o'clock.

Society of Instrument Technology

27th November. "Television Technique Applied to Observation and Control", by Professor J. D. McGee, O.B.E., M.Sc., Ph.D., at 7 o'clock at Manson House, Portland Place, London, W.1.

Radar Association

14th November. "Infra-Red: Its Problems and Possibilities", by Dr. F. E. Jones, M.B.E., B.Sc., Ph.D., under the chairmanship of Sir Robert Renwick, Bt., K.B.E., President of the Association, to be held at 7.30 at the Anatomy Theatre, University College, Gower Street, London, W.C.1. (Open to non-members.)

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

ALTERATION TO MSF 60 kc/s

Due to extensive maintenance work on the long-wave aerial system at Rugby, it will no longer be possible, after the end of October, to transmit MSF 60 kc/s at 1429-1530 G.M.T. In order to maintain a service to users, an alternative transmission has been arranged and for a period of about eighteen months, commencing on the 1st November 1956, MSF 60 kc/s will be radiated at 1959-2100 G.M.T. The values of the transmitted frequency will continue to be published in *Wireless Engineer*.

Values for September 1956

Date 1956 September	MSF 60 kc/s Frequency deviation from nominal*: parts in 10^9
1	+1
2	+1
3	0
4	+1
5	+1
6	+2
7	+2
8	N.M.
9	+2
10	+1
11	+2
12	+2
13	+2
14	+1
15	+1
16	+2
17	+2
18	+1
19	+1
20	+1
21	+1
22	+1
23	+1
24	+1
25	0
26	+1
27	+1
28	+2
29	+2
30	+2

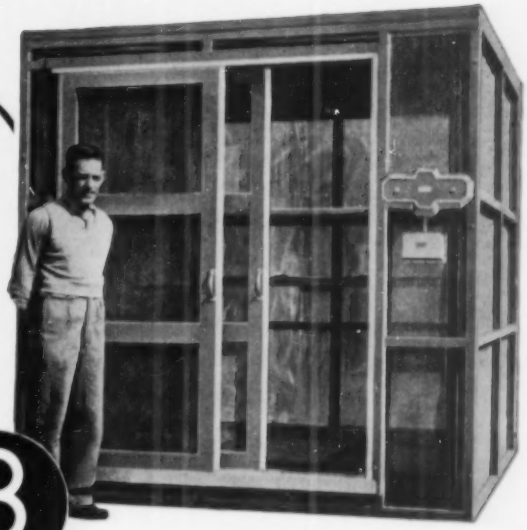
N.M. = Not Measured.

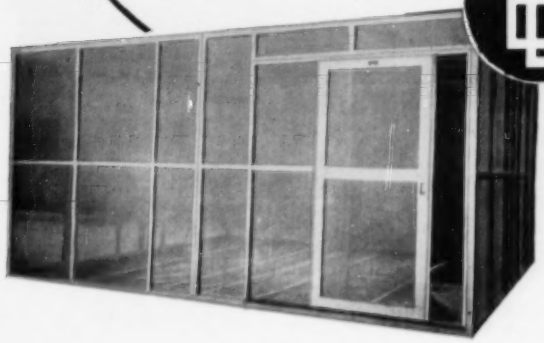
*Nominal frequency is defined to be that frequency corresponding to a value of 9 192 631 830 c/s for the N.P.L. caesium resonator.

SCREENED ROOMS

(FARADAY CAGES)

Two typical screened compartments designed to customers' special requirements. Designs and prices submitted on request.





A screened compartment may be used in two ways: as a barrier surrounding a source of radio frequency energy (as from electro-medical or R.F. heating processes) to prevent interference with radio, or to exclude R.F. energy, e.g. as a "quiet room" for the alignment and test of sensitive electronic apparatus.

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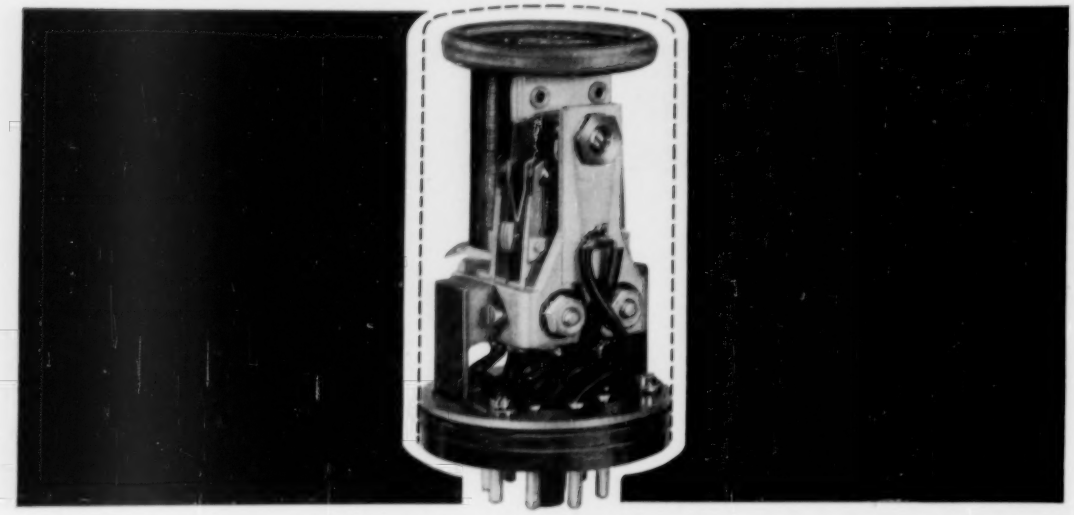

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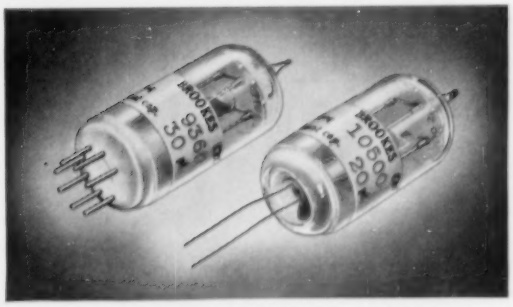

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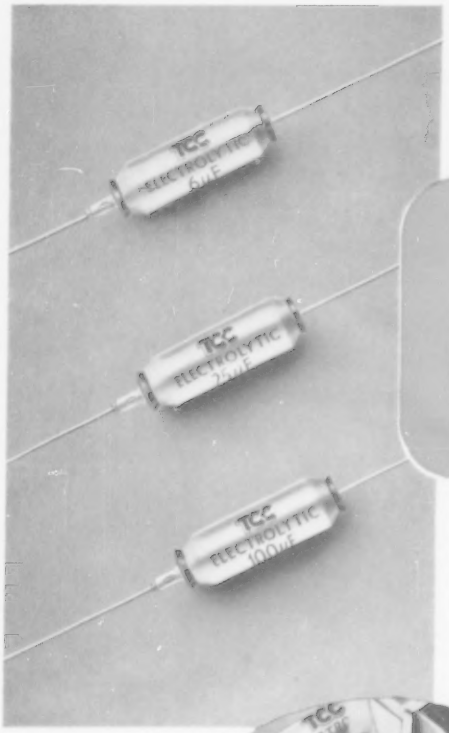

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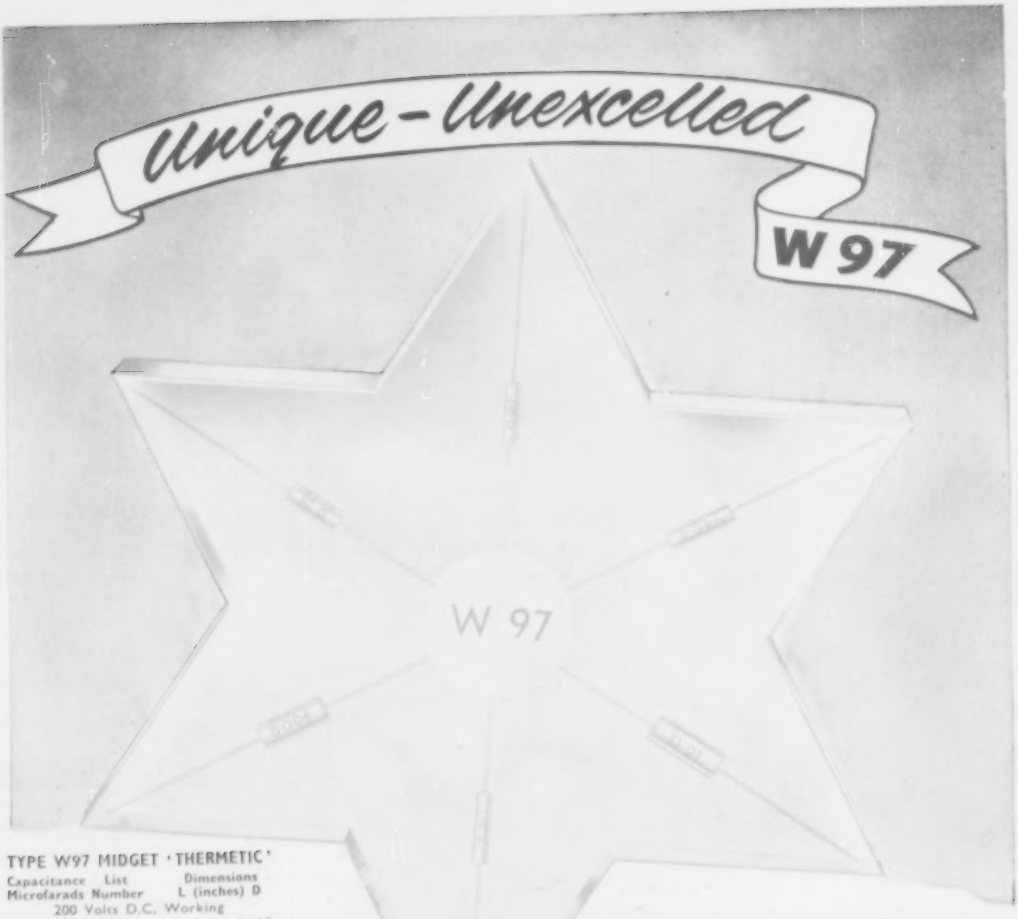

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0.005	BM12	0.500	0.180
0.01	BM13	0.500	0.180
0.02	BM14	0.610	0.180
0.03	BM15	0.610	0.260
0.04	BM16	0.610	0.260
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0.001	BM6	0.610	0.135
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0.01	BM21	0.610	0.260
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0.00022	BM28	0.500	0.180
0.00025	BM29	0.500	0.180
0.0003	BM3	0.610	0.135
0.0003	BM30	0.500	0.180
0.0004	BM36	0.500	0.180
0.0005	BM31	0.500	0.180
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